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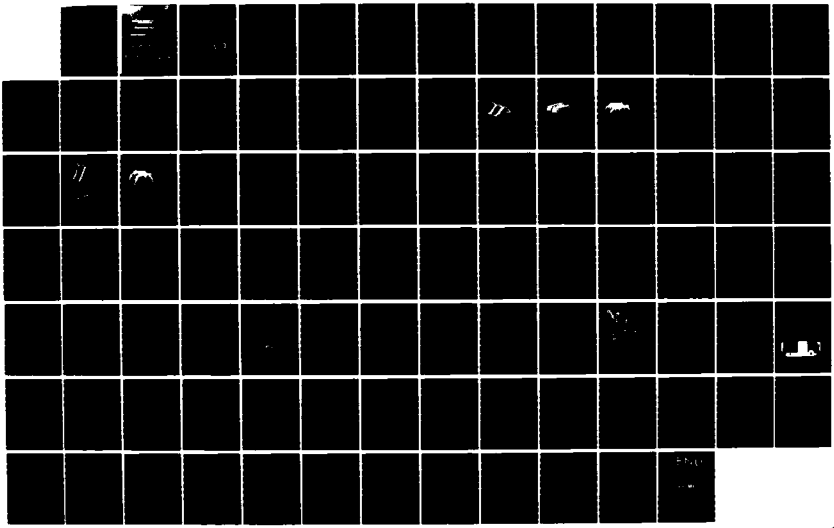

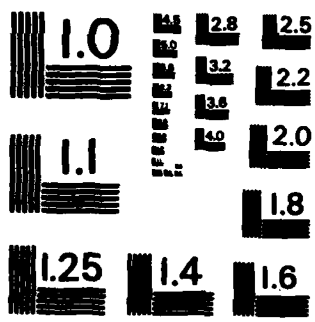

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Analysis of a currently used low profile train antenna. Calculation of the antenna losses, coupler losses, overall efficiency and radiation patterns. Development of an improved antenna structure and calculation of the corresponding parameters. Experimental verification of the theoretical results.

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T a b l e   o f   C o n t e n t s

1. Introduction

For telecommunication links in the HF range an antenna installed on top of a railroad train had to be optimized. The specifications are:

- total maximum height of antenna: h_{max} = 0.23m
- total maximum length of antenna: a_{max} = 25m
- tunable to a source impedance of 50 Ohm by help of a standard type antenna coupler in the frequency range from 3 to 5MHz and a maximum standing wave ratio of SWR_{max} = 2.

Because of the height limitations only a loop antenna is usable. One train, equipped with a loop antenna of 7.5m length, 0.23m height and 0.075m diameter (Fig.1) has been investigated. This antenna will be referred to as "type 1".

Another antenna construction is proposed in /1/. It consists of two symmetrical antenna parts fed by a coaxial line, the outer conductor of which is one antenna element (Fig.2). This shielded-loop antenna will be referred to as "type 2".

The two antenna types differ in

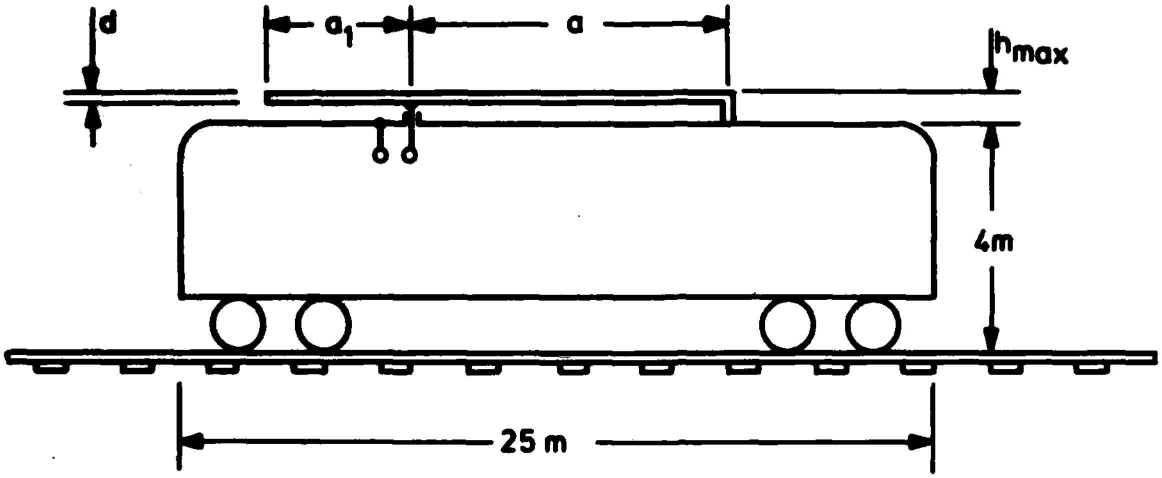
- impedance (real- and imaginary part)
- radiation pattern.

The radiation resistance of both antenna types is very low, but the resistance of type 1 is significantly lower than type 2. The imaginary part of the antenna impedance is inductive for type 1 and capacitive for type 2, if the length is optimized. Therefore type 2 can be matched easier to the source impedance with a coupler that has been constructed for a capacitive whip antenna (Harris antenna-coupler RF-615B).

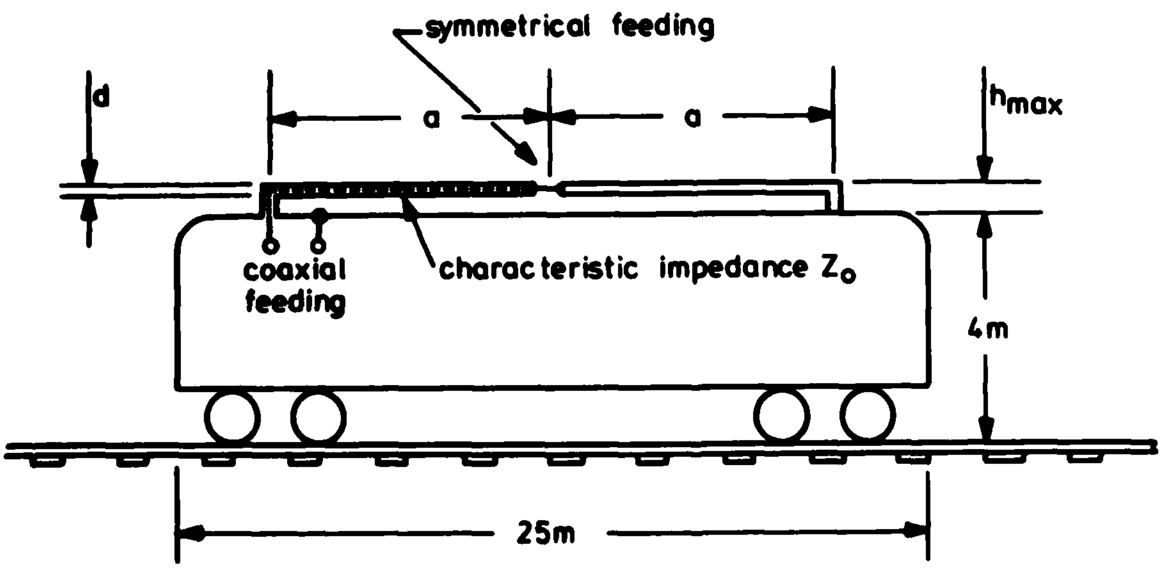
The radiation pattern in the horizontal plane is a figure-eight pattern for both antennas. For ground-wave links, type 2 will only work sufficiently in the direction the train goes to or comes from. Type 1 radiates omnidirectional since the the opposite fields of the two vertical antenna elements (i.e. feeder and short circuit) have different magnitudes.

The main problem of the total antenna system is the coupler. Several Milliohms of radiation resistance has to be transformed to 50 Ohms, while some hundred Ohms of the imaginary part of the antenna impedance has to be compensated for. Typically about 90 percent of the transmitter's available power is converted to heat inside the coupler.

According to the strong influence of the coupler on the whole



<u>Fig. 1:</u> Outline dimensions of antenna type 1:
$a = 7.5m$, $a_1 = 1.5m$, $d = 0.075m$, $h_{max} = 0.23m$



<u>Fig. 2:</u> Construction of antenna type 2

antenna system it is obviously necessary to know precise data of all coupler elements. This data was taken from a slightly smaller, but comparable coupler.

With this data the total antenna efficiency ((radiated power)/(power, available from the generator)) for type 1 has been calculated utilizing the length, the diameter and the number of antenna elements and for type 2, the influence of the characteristic impedance of the coaxial feeder-line.

It can be shown that at the most critical frequency - 3MHz - a total efficiency of up to 12% is possible if the optimum value of each parameter is chosen. On the other hand the antenna optimized for 3MHz cannot be matched to the transmitter at 4MHz or 5MHz with the available HARRIS-coupler. Because of this, a compromise between maximum efficiency and necessary tuning bandwidth must be made.

2. Antenna calculation

The dimensions of antenna type 1 and 2 are shown in Fig.1 and Fig.2, respectively.

To simplify the numerical calculations, the conductivity of the antenna elements is initially assumed to be infinite. Therefore, the cylindrical surface of every antenna element can be substituted by a line current at a height of

$$h = \sqrt{h_{max} \cdot (h_{max} - d)} \qquad (1)$$

above ground. The real antenna surface coincides with one equipotential surface of this current filament (see Fig.3).

The roof of the train can be substituted by using the image of the real antenna.

The distance between roof and ground can be neglected, as the current density at the vertical walls of the train is definitely much smaller than the current density at the antenna elements and their images (as in reality, the antenna is close to the roof).

Because of these simplifications the final configurations of the antennas are as shown in Fig.4.

2.1 Radiation pattern

The radiation pattern can be calculated by summing-up the retarded potentials of the antenna currents far away from the antenna (far-field pattern) for each direction. In /1/ this has

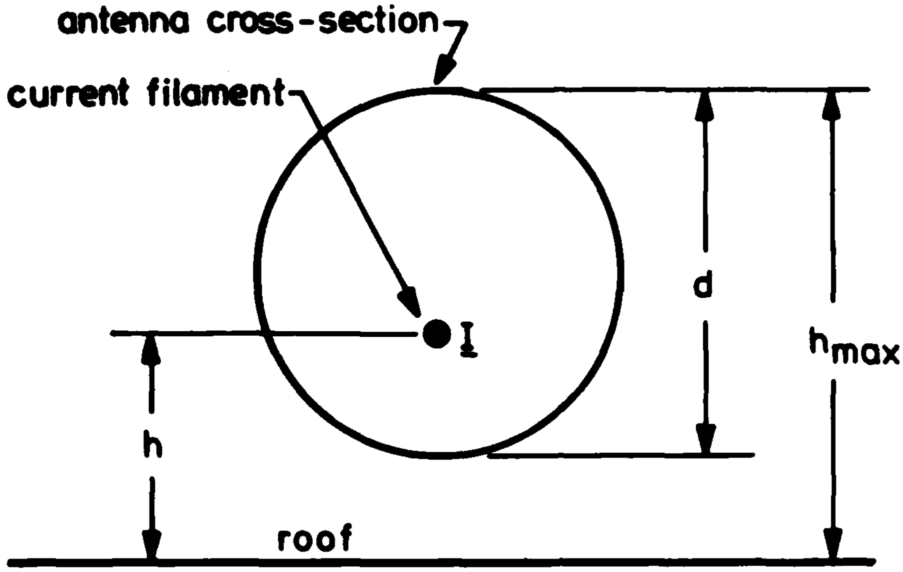


Fig. 3a: Substitution of the conducting antenna surface by a current filament $\underline{I}$ of the height h above ground (roof)

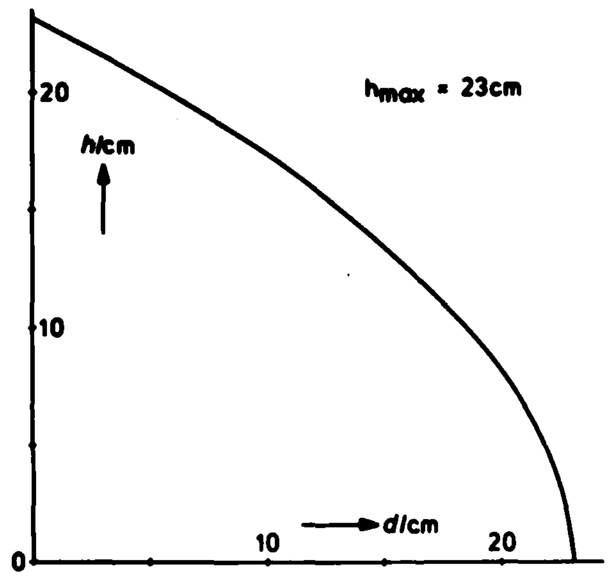


Fig. 3b: Hight h of the current filament versus the antenna diameter d

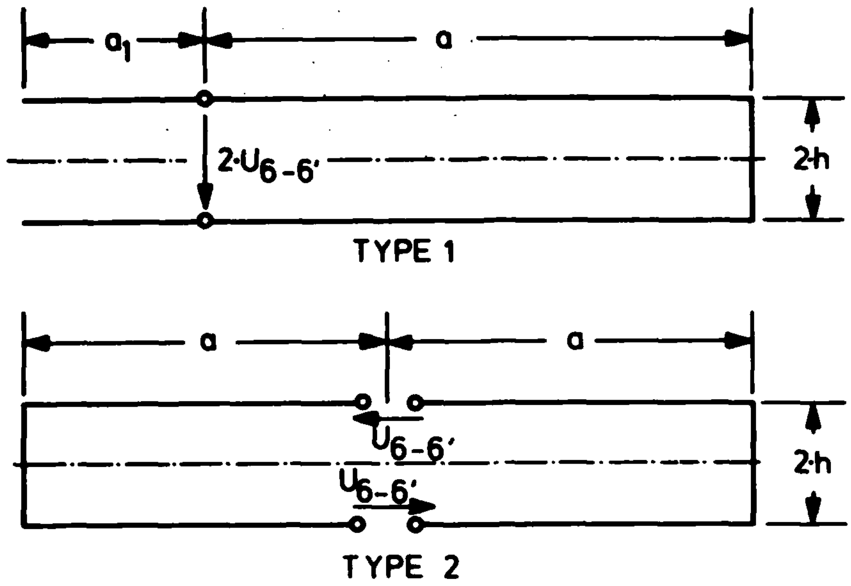


Fig. 4: Simplified configuration of antenna type 1 and type 2

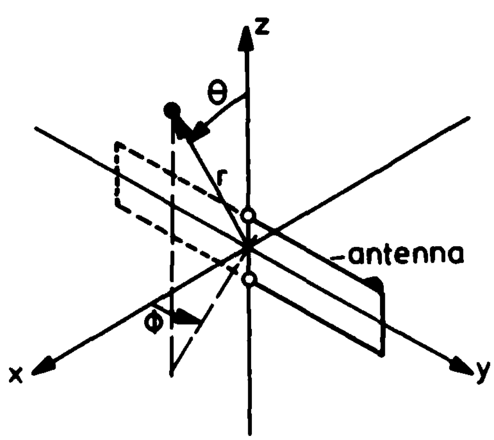


Fig. 5: Position of the antenna in the centre of the used coordinates

been done in a closed analytic form.

To enable a more simplified explanation a numerical method has also been used:

As shown in Fig.6 the simplified antenna is positioned to the centre of a spherical (ϕ,θ,r) coordinate system. In a corresponding x-y-z-system the railroad would coincide with the y-axis. The antenna current in a good approach is sinusoidal, with a maximum current through the shortcircuit termination, represented by the vertical rods $(y = \pm a)$:

$$\underline{I}(y) = \underline{I}(o)\cdot\frac{\cos[\beta_o(a-|y|)]}{\cos(\beta_o\cdot a)} \tag{2}$$

where $\beta_o = \frac{2\pi}{\lambda_o}$

and I(0) = current at the feeding point.

The antenna and its image are then subdivided into k parts, each with approximately constant current distribution.

Each part acts as a Hertzian electric dipole of the length l:

$$l = \frac{a}{k+1} \tag{3}$$

The magnetic far-field at the distance R_o in the directions ϕ and θ (vector $\vec{r}_o$), caused by the i-th horizontal electric current element $\underline{I}_i = \underline{I}(y = y_i)$ is:

$$\underline{H}_\theta = j\cdot\frac{\underline{I}_i}{2\cdot R_o}\cdot\frac{l}{\lambda_o}\cdot\cos\phi \tag{4}$$

$$\underline{H}_\phi = -j\cdot\frac{\underline{I}_i}{2\cdot R_o}\cdot\frac{l}{\lambda_o}\cdot\cos\theta\cdot\sin\phi \tag{5}$$

The locus vector $\vec{r}_i$ of the centre of each horizontal Hertzian electric dipole is in Cartesian coordinates:

$$\vec{r}_i = \begin{pmatrix} 0 \\ \frac{l}{2} + i\cdot l \\ \pm h \end{pmatrix} \tag{6}$$

where i = 0 ... k for type 1 and i = -k ... k for type 2

The path difference of a wave caused by the i-th electric current element with respect to the antenna centre is:

$$\Delta_i = \vec{r}_i\cdot\frac{\vec{r}_o}{r_o} \tag{7}$$

For any current element with the coordinates x_i, y_i, and z_i this leads to:

$$\Delta_i = x_i \cdot \sin\theta \cdot \cos\phi + y_i \cdot \sin\theta \cdot \sin\phi + z_i \cdot \cos\theta \tag{7a}$$

The locus vector of the centers of either the type 1 or the type 2 vertical current elements is:

$$\vec{r}_i = \begin{pmatrix} o \\ \pm a \\ o \end{pmatrix} \tag{8}$$

The current in this element is the maximum current of the antenna

$$\underline{I}_m = \frac{\underline{I}(o)}{\cos(\beta_o \cdot a)} \tag{9}$$

The element's length is:

$$l_m = 2 \cdot h \tag{10}$$

and its magnetic far-field is:

$$\underline{H}_\phi = -j \cdot \frac{\underline{I}_m}{2 \cdot R_o} \cdot \frac{l_m}{\lambda_o} \cdot \sin\theta \tag{11}$$

Antenna type 1 is usually fed through the train's roof. The feeder line with the length $l_m = 2h$ causes the magnetic far-field:

$$\underline{H}_\phi = j \cdot \frac{\underline{I}(o)}{2 \cdot R_o} \cdot \frac{l_m}{\lambda_o} \cdot \sin\theta \tag{11a}$$

if the length of the open-ended parallel line is $a_1 = 0$. Otherwise the feeder current is smaller than the horizontal current I(0).

For any point P with the locus vector $\vec{r}_o$ the field strengths become:

$$\underline{H}_\theta = j \cdot \frac{l \cdot \cos\phi}{2 \cdot R_o \cdot \lambda_o} \cdot \sum_{i=s}^{i=k} \underline{I}_i \cdot [\exp(-j \cdot \beta_o \cdot \Delta_i) - \exp(j \cdot \beta_o \cdot \Delta_i)] \tag{12}$$

$$\underline{H}_\phi = -j\cdot\frac{l\cdot\cos\Theta\cdot\sin\phi}{2\cdot R_o\cdot\lambda_o}\cdot\sum_{i=s}^{i=k}\underline{I}_i[\exp(-j\cdot\beta_o\cdot\Delta_i)-\exp(j\cdot\beta_o\cdot\Delta_i)]\ -$$

$$-\ j\,\frac{\underline{I}_m\cdot l_m\cdot\sin\Theta}{2\cdot R_o\cdot\lambda_o}\cdot\sum_{i=1}^{i=m}\exp(-j\cdot\beta_o\cdot\Delta_i)+j(2-m)\cdot\underline{I}(o)\cdot\frac{l_m\cdot\sin\Theta}{2\cdot R_o\cdot\lambda_o} \qquad (13)$$

where s = 0 and m = 1 for type 1 and s = -k and m = 2 for type 2.

The total magnetic field strength is

$$H = \sqrt{|\underline{H}_\phi|^2 + |\underline{H}_\theta|^2} \qquad (14)$$

The total electric field strength is

$$E = H\cdot Z_i \qquad (15)$$

(Intrinsic impedance Z_i = 120πΩ)

The average Poynting vector is

$$S = \frac{1}{2}\cdot E\cdot H$$

$$= \frac{1}{2}\cdot\frac{E^2}{Z_i} \qquad (16)$$

The radiation patterns of the two antenna types for ϕ and θ-polarization and the total field strength, respectively, have been calculated and mapped.

Fig.6 to Fig.8 show the calculated field strengths for antenna type 1. For this pattern, as well as for the following, the current I(0) at the antenna's feeding point is assumed to be constant. The patterns in the ground plane (Θ = 90°) would be exactly circular, if no vertical feeder current would be necessary. Because of the feeder current, which has a 180° - phase shift to the current in the short circuit the radiation in the directions ϕ = 0° and ϕ = 180° is smaller than in the directions ϕ = ±90°.

The field strength increases with the frequency, as the maximum current I_m increases depending on eq.9.

The radiation patterns in the plane ϕ = 90° is shown in Fig.7. In this plane the radiation is nearly independent from the elevation angle.

The radiation patterns in the plane ϕ = 0° show that radiation

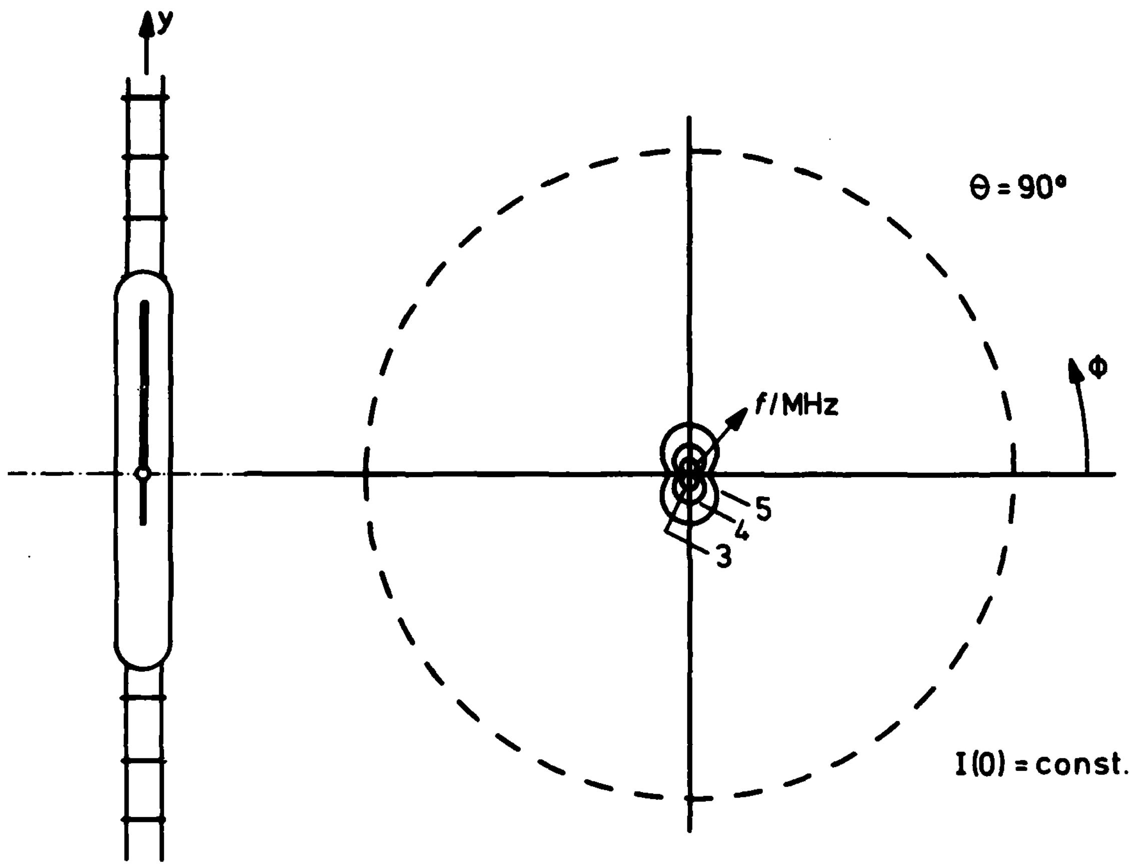


Fig. 6: Antenna type 1 - Total electrical field strength in the ground plane. Current I(o) at the feeder point is constant

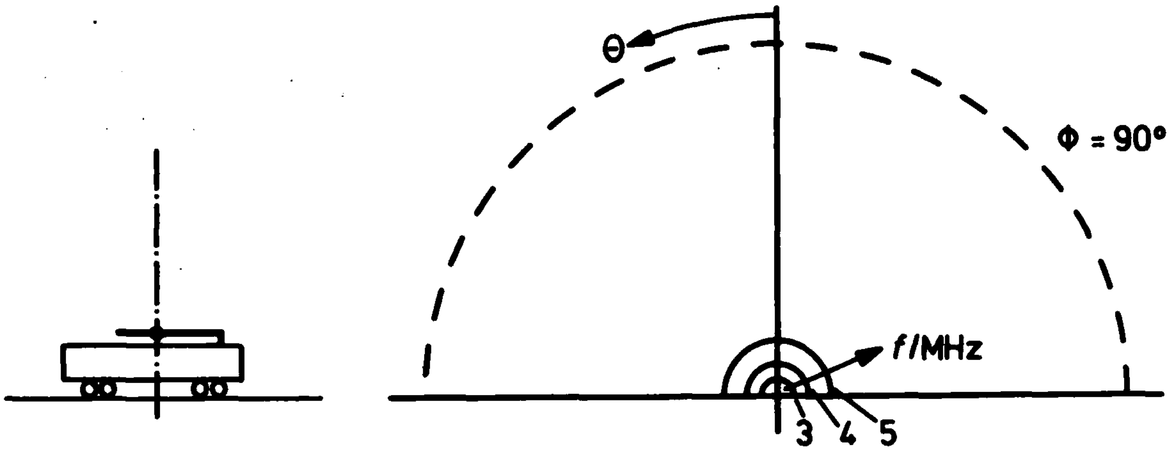


Fig. 7: Antenna type 1 - Total electrical field strength in the plane $\phi = 90^{o}$. I(o) = const.

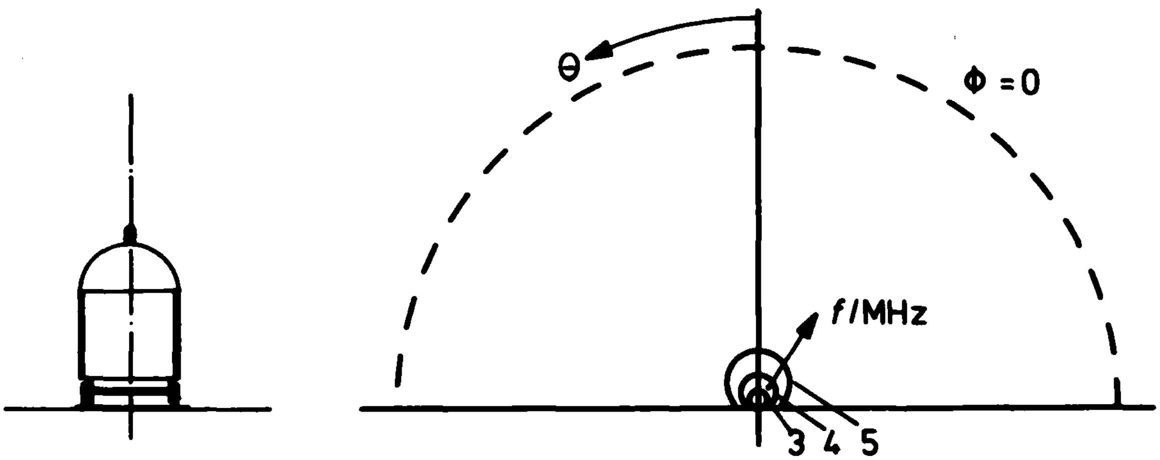


Fig. 8: Antenna type 1 - Total electrical field strength in the plane $\phi = 0^{o}$. I(o) = const.

minima occur only at lower elevation angles (Fig.8).

Antenna type 2 has two vertical elements spaced apart the distance 2a with opposite currents I_m. The fields of these two elements, therefore, are cancelled in the plane $\phi = 0^\circ$. As shown in Fig.9 ground-wave propagation does not occur perpendicular to the train's direction. Under most conditions this will not be disadvantageous.

Sky-wave communication links are possible in any direction, as shown in Fig.10 and Fig.11. The radiation to higher angles for type 2 is at 4MHz about 14dB higher than with antenna type 1 if the same current I(0) is flowing. It just has been pointed out that the transmitter power predominantly heats the coupler. Because of this, absolute values of the field strength can only be compared with constant available generator power. As this will be done later, and as it does not change the relative radiation pattern, the following patterns are related to the maximum field strength of each antenna at the particular frequency.

As the normalized radiation patterns obviously do not depend strongly on the frequency, only the patterns for one frequency (3MHz) are presented in the different charts.

Fig.12 shows the partial electrical field strength with ϕ-polarization of antenna type 1 in the plane $\phi = 0^\circ$. In the planes $\phi = 90^\circ$ and $\theta = 90^\circ$ this partial field strength is zero. The θ-polarized electrical field strength is represented in Fig.13 and the total fieldstrength in Fig.14, both of them in the planes $\phi = 90^\circ$, $\phi = 0^\circ$ and $\theta = 90^\circ$. This shows that in the ground plane only vertical polarization occurs while at higher elevation angles ϕ-polarization will be found as well as θ-polarization.

Fig. 15 to Fig. 17 show the radiation patterns of antenna type 2, again in the planes $\phi = 0^\circ$, $\phi = 90^\circ$ and $\theta = 90^\circ$. The ϕ-polarized E-field strength occurs, similar as with type 1, only in the plane $\phi = 0^\circ$. The θ-polarized E-field strength is plotted in Fig.16. It has only slight changes in the plane $\theta = 90^\circ$ with a maximum at $\theta = 0^\circ$, but has a zero at $\phi = 0^\circ$ and $\theta = 90^\circ$. The polarization in the ground plane is vertical as with type 1, as the ground conductivity is assumed to be infinite. The total field strength in the upper half space is plotted for the planes $\phi = 0^\circ$, $\phi = 90^\circ$ and $\theta = 90^\circ$ in Fig.17. This illustrates that the radiation pattern of this antenna is half a toroid with its axis in the ground plane.

Fig.18 to Fig.20 show for antenna type 1 the dependence of the ϕ- and θ-polarized electric field strength as a three-dimensional plot over the rectangular ϕ-θ-plane. Using this chart, the relative field strength for every ϕ-θ-direction of the upper half

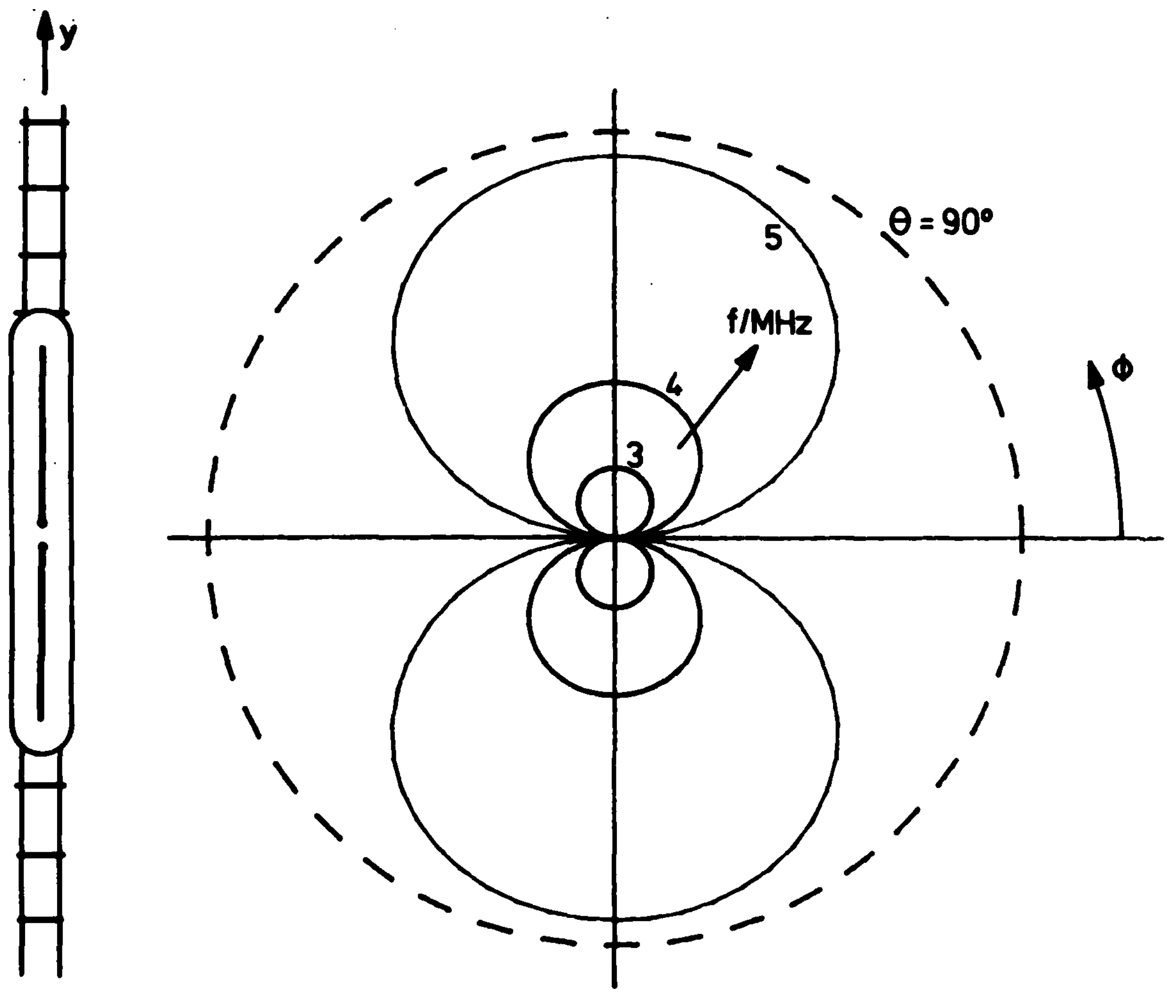


Fig. 9: Antenna type 2 - Total electrical field strength in the ground plane. Current I(o) at the feeding point is constant.

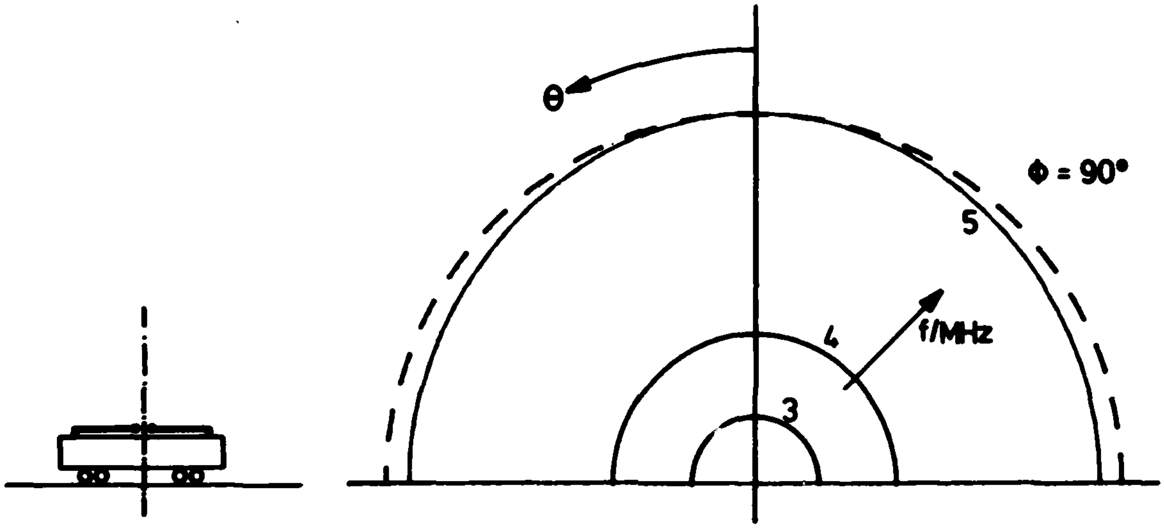


Fig. 10: Antenna type 2 - Total electrical field strength in the plane $\phi = 90^{o}$. I(o) = const.

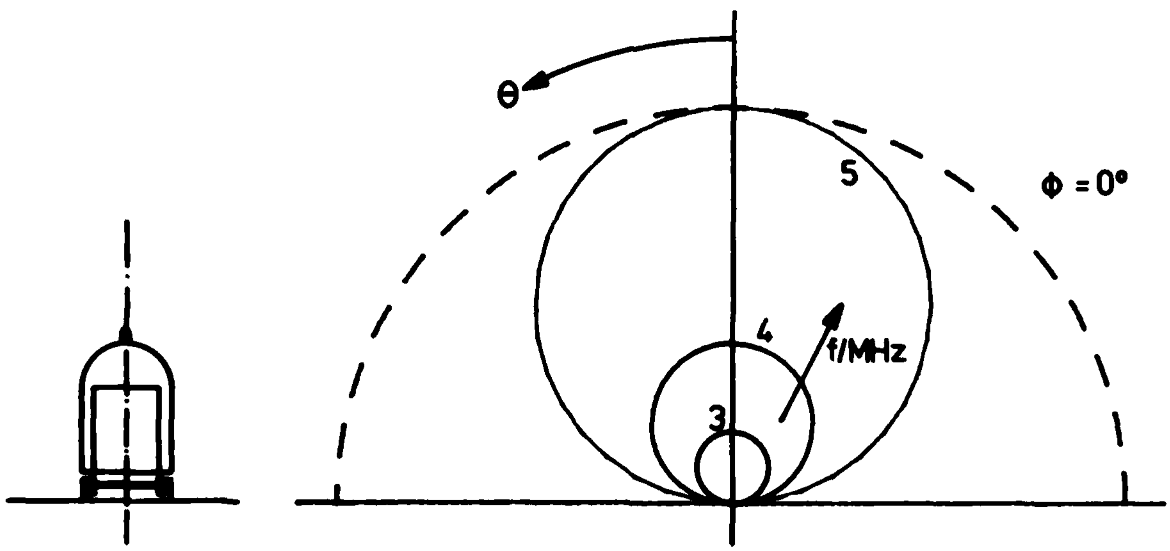


Fig. 11: Antenna type 2 - Total electrical field strength in the plane $\phi = 0^{o}$. I(o) = const.

TYPE 1

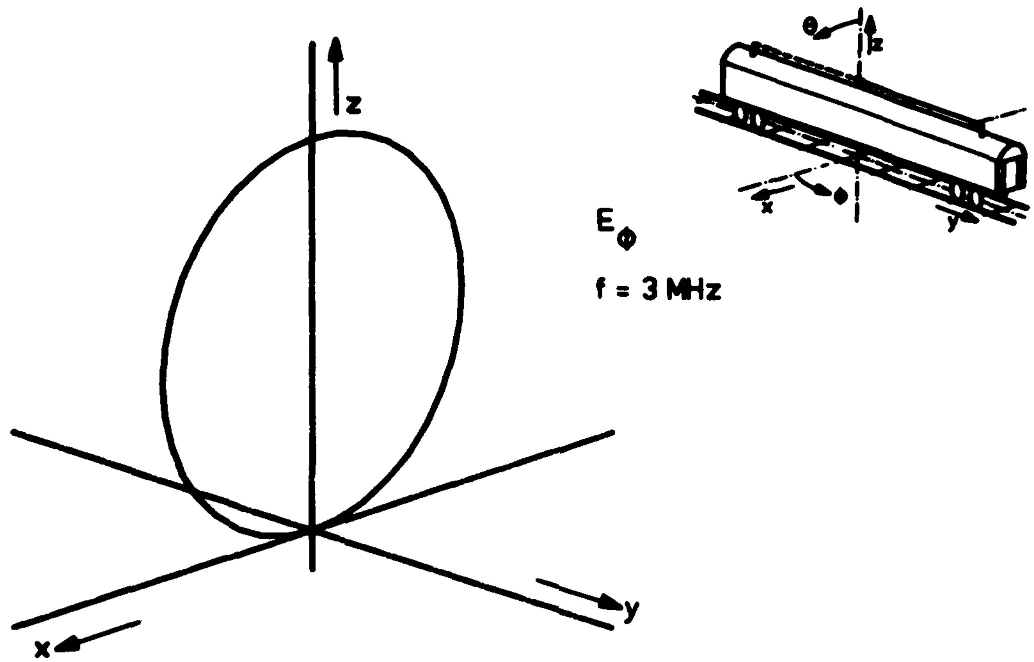


Fig. 12: Partial ϕ-polarized electrical field strength

TYPE 1

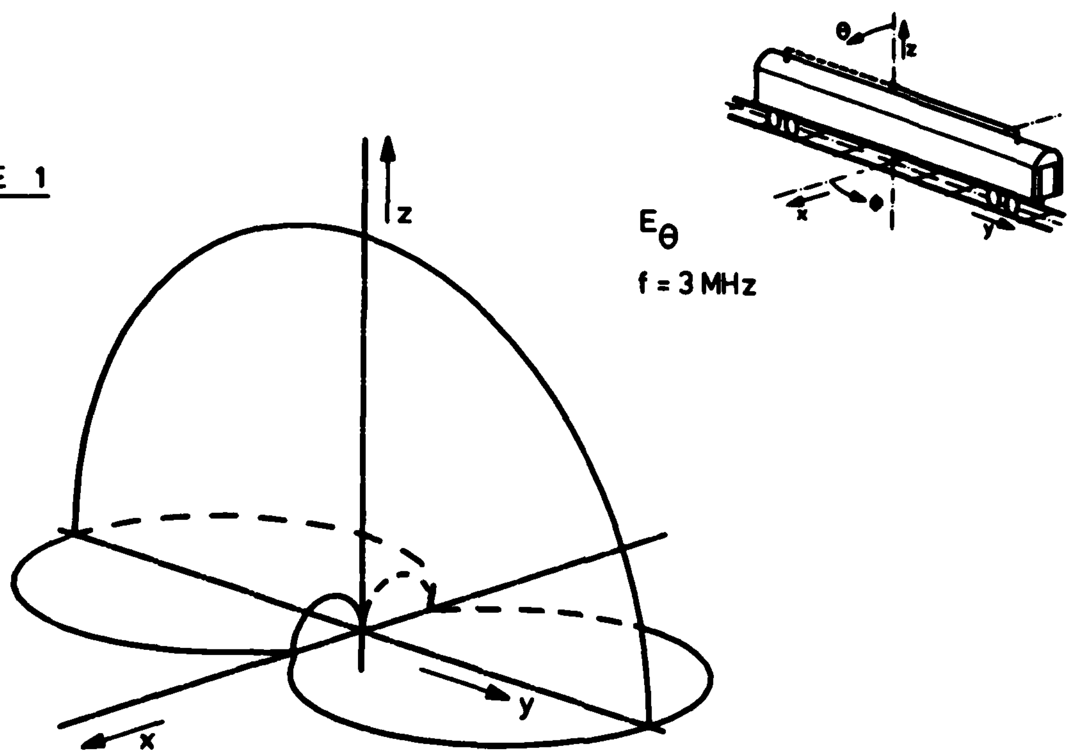


Fig. 13: Partial Θ-polarized electrical field strength

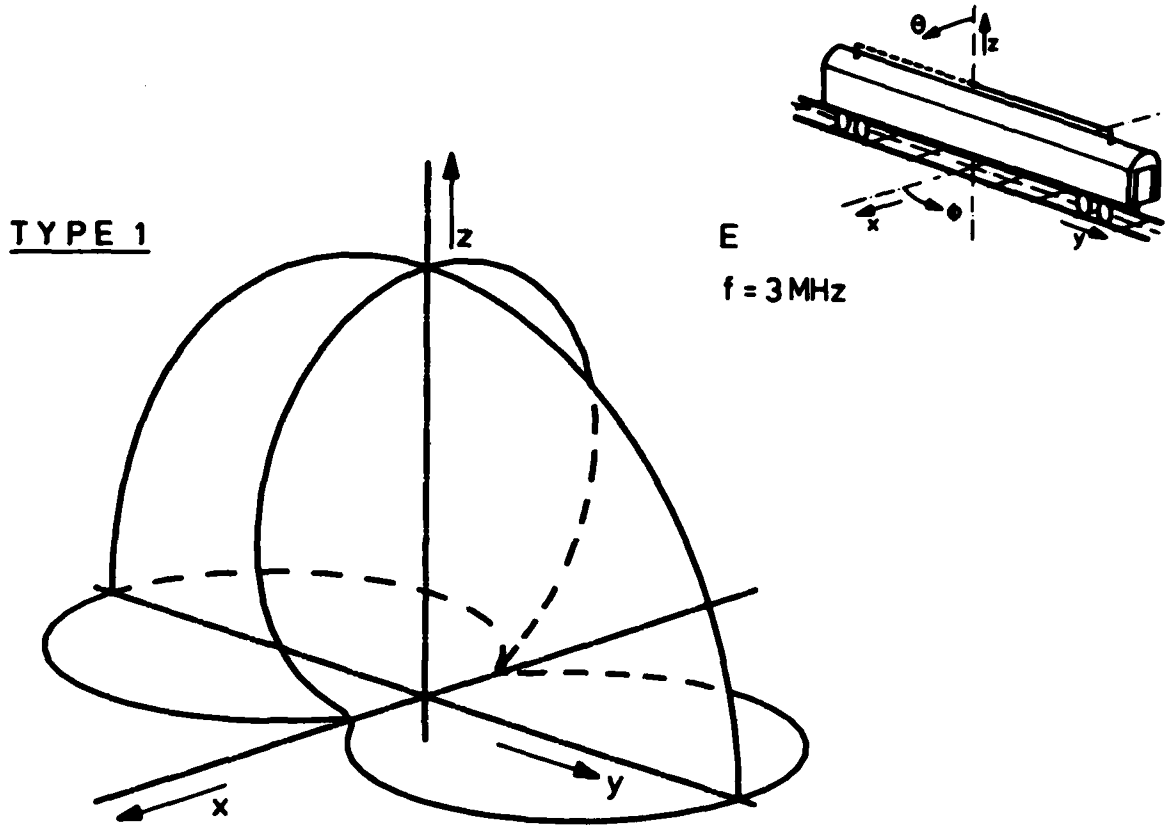


Fig. 14: Total electrical field strength

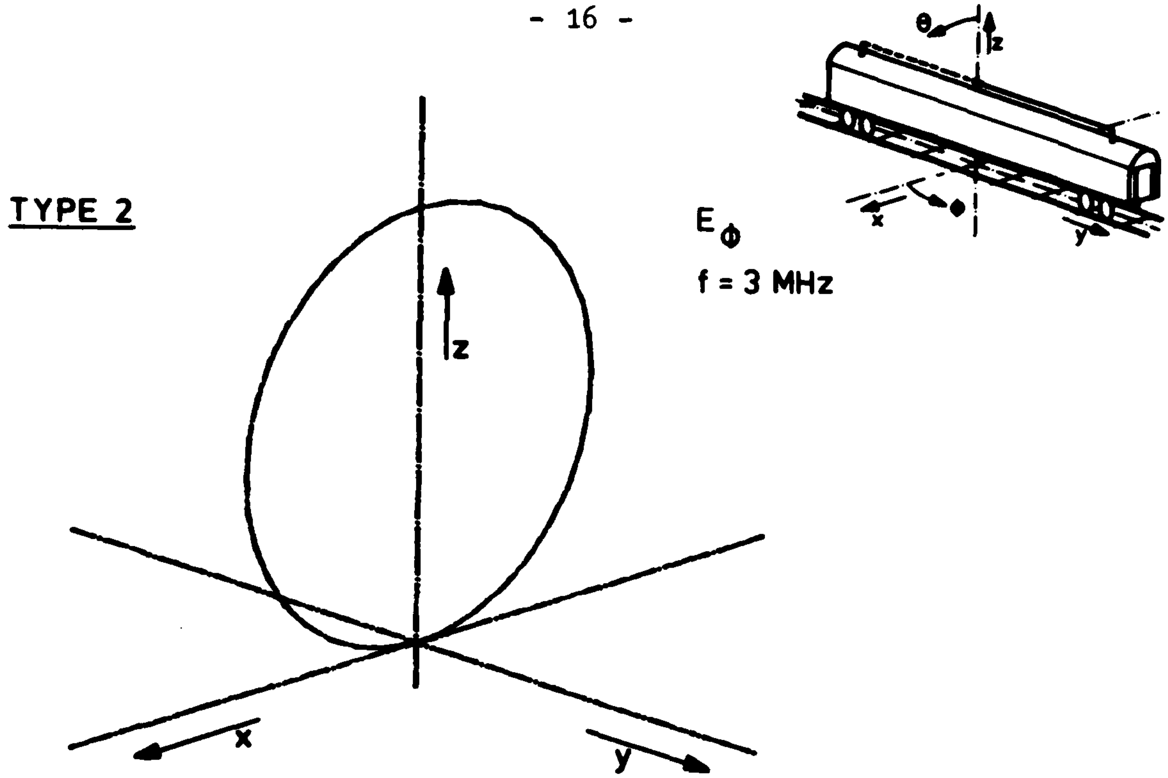


Fig. 15: Partial ϕ-polarized electrical field strength

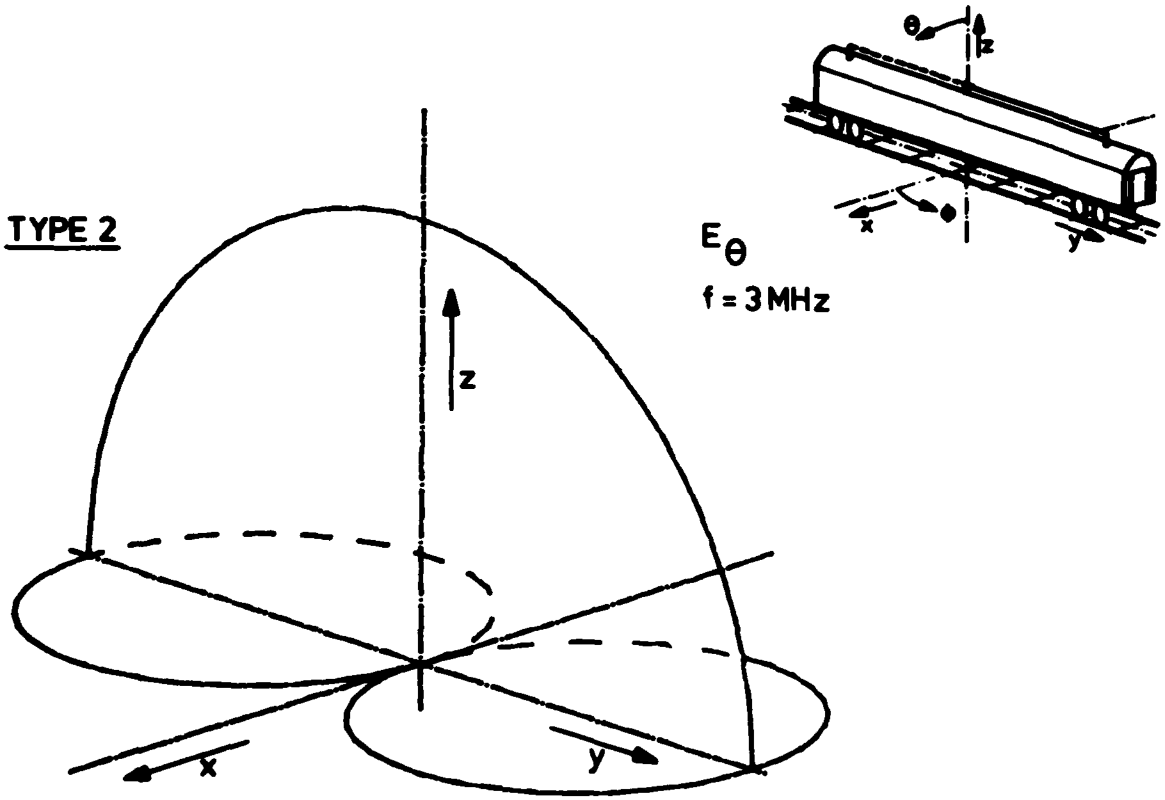


Fig. 16: Partial Θ-polarized electrical field strength

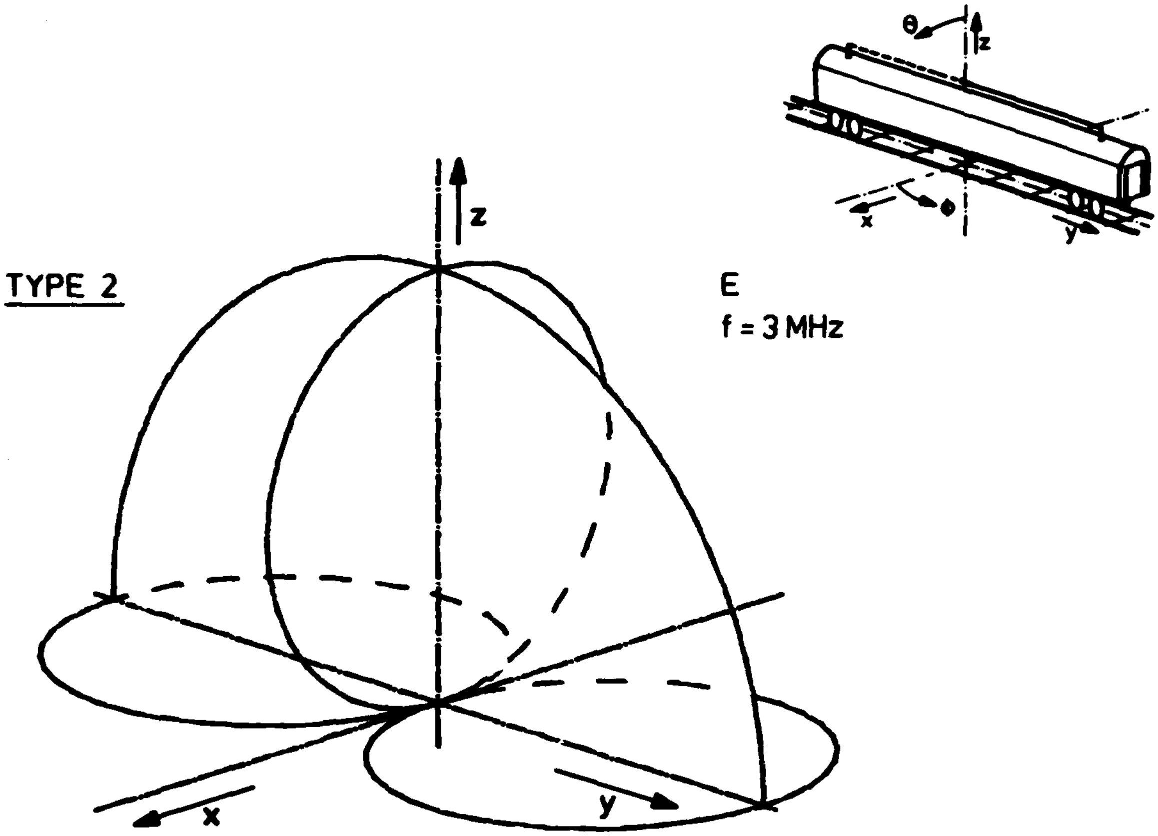


Fig. 17: Antenna type 2 - Total electrical field strength

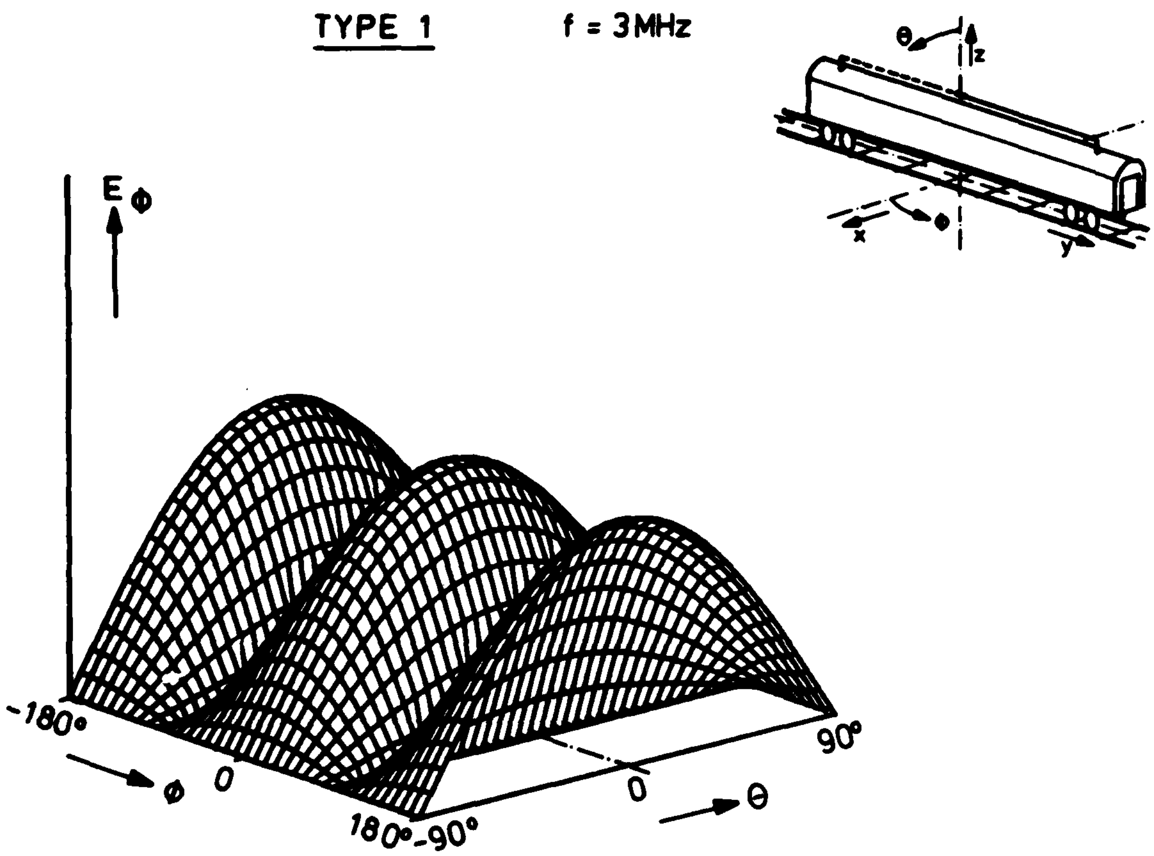


Fig. 18: Antenna type 1 - partial ϕ-polarized electrical field strength over the rectangular ϕ-Θ-plane

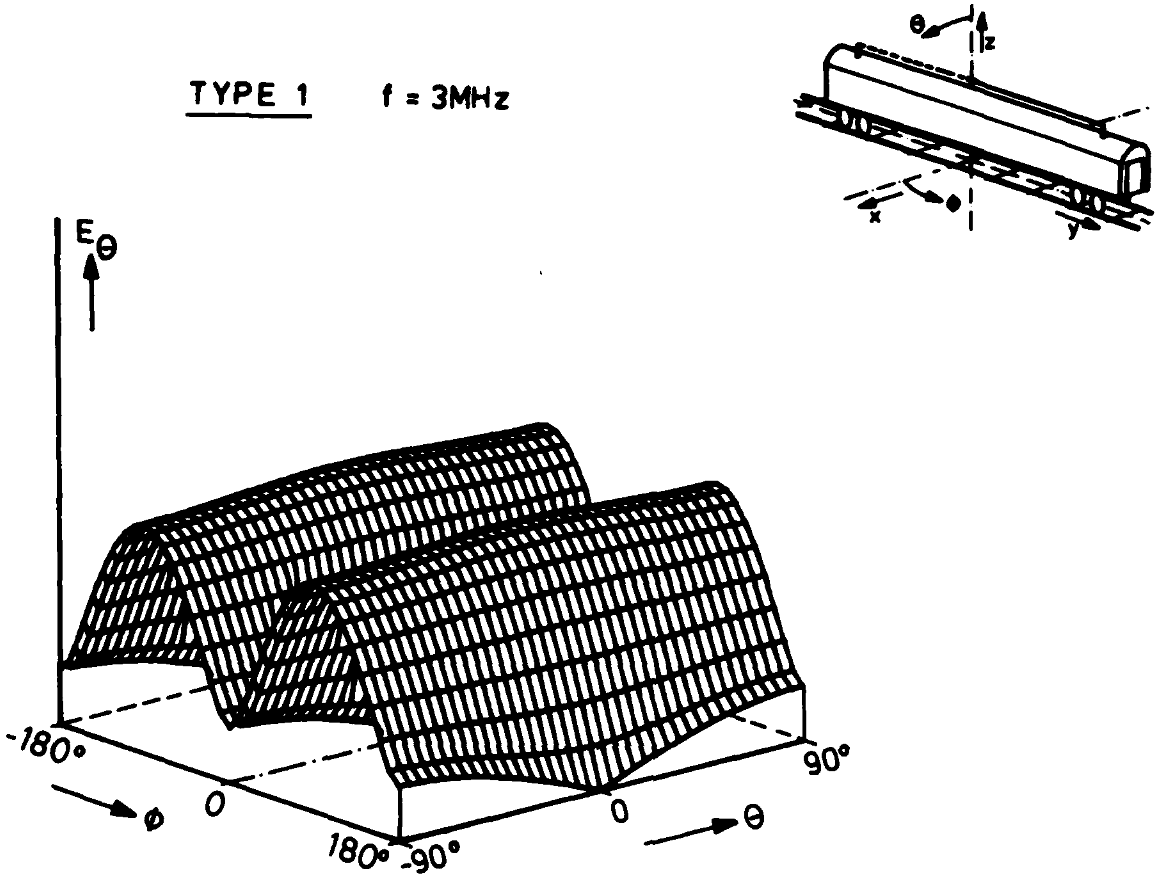
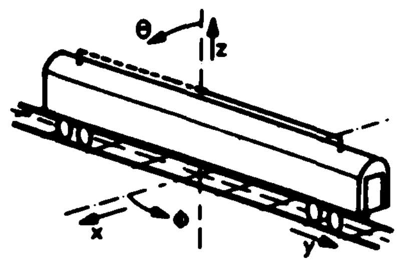


Fig. 19: Antenna type 1 - partial Θ-polarized electrical field strength over the rectangular ϕ-Θ-plane

TYPE 1 f = 3MHz



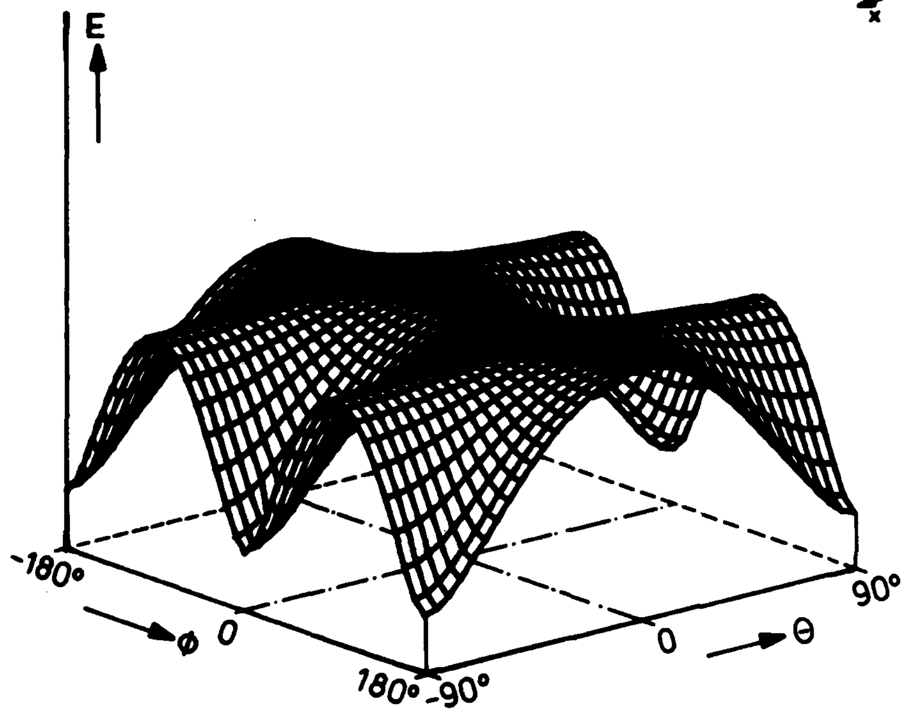


Fig. 20: Antenna type 1 - Total electrical field strength over the rectangular ϕ-Θ-plane

space can be obtained. Therefore, the maximum and minimum values can be found easily. Maximum radiation of antenna type 1 with ϕ-polarization is at $\Theta = 0^{\circ}$ and $\phi = 0^{\circ}$ and $\phi = \pm 180^{\circ}$ respectively. No radiation occurs in those directions where $\phi = \pm 90^{\circ}$ or where $\Theta = \pm 90^{\circ}$. The maximum of Θ-polarized radiation is found at $\Theta = \pm 90^{\circ}$ and $\theta = \pm 90^{\circ}$. No radiation exists at $\Theta = 0^{\circ}$ and $\phi = 0^{\circ}$ or $\phi = \pm 180^{\circ}$.

The total field strength pattern has a minimum at $\phi = 0^{\circ}$ and $\phi = \pm 180$, when $\Theta = \pm 90^{\circ}$. Its maximum is at $\phi = \pm 90^{\circ}$, and nearly independent of Θ.

Fig.21 to Fig.23 represent the same functions like Fig.18 to Fig.20, but are plotted as constant-field strength curves over the rectangular ϕ-Θ-plane.

The radiation patterns of antenna type 2 over the rectangular ϕ-Θ-plane, represented in a three-dimensional and a constant-field strength plot, respectively, are shown in Fig.24 to Fig.29.

The ϕ-polarized electric field strength (Fig.24 and Fig.27) of this antenna is distributed similar to that of antenna type 1. The Θ-polarized field shows a maximum value for $\phi = \pm 90^{\circ}$, which is nearly independent of the elevation. Zeros exist for $\phi = 0^{\circ}$, 180° and 360°.

The total electrical field strength is found to be zero for $\Theta = \pm 90^{\circ}$ and $\phi = 0^{\circ}$, 180° and 360°. On this account wave propagation normal to the driving direction is not possible for both polarizations, if the elevation angle θ is exactly $\Theta = 90^{\circ}$. Therefore, in some unusual cases ground-wave communications might fail in a small angle range, while sky-wave communications is possible in any direction.

2.2 Radiation resistance

The power radiated from the antenna into the space can be calculated by integrating the power density over the radiation sphere:

$$P_r = \int S \cdot da \tag{17}$$

where:

$$da = R_o \cdot \sin\Theta \cdot d\Theta \cdot d\phi \tag{18}$$

As the antenna is initially assumed to be lossless, this radiated power has to be identical with the antenna input power:

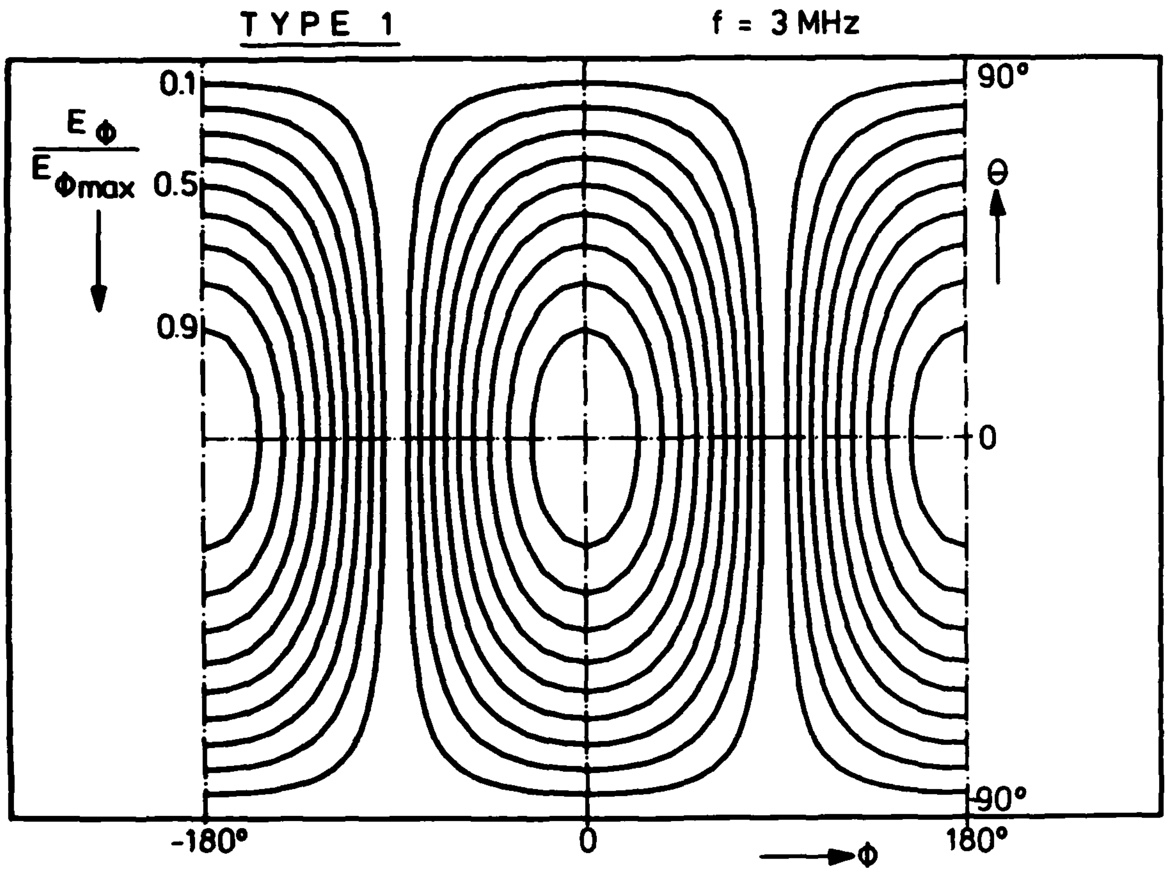


Fig. 21: Antenna type 1 - lines of constant electrical field strength over the rectangular ϕ-Θ-plane (ϕ-polarization)

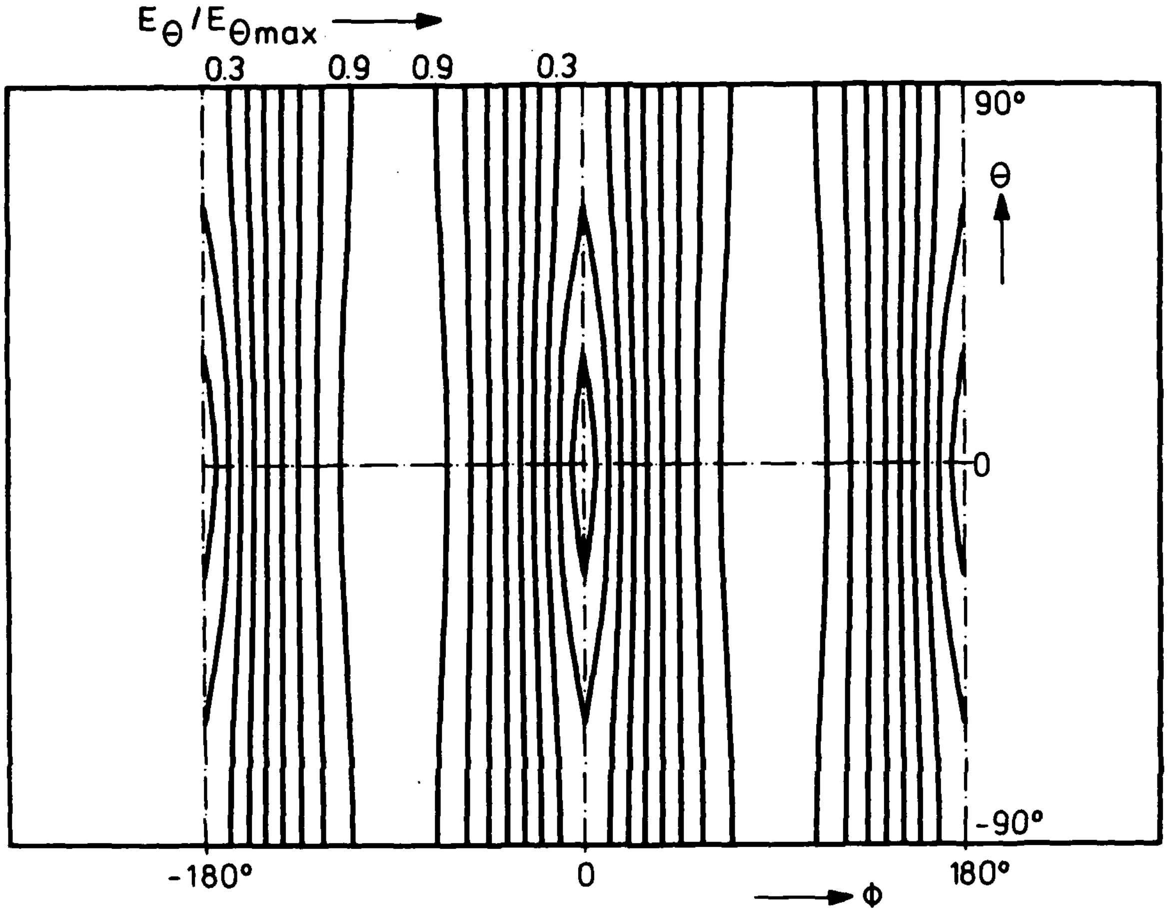


Fig. 22: Antenna type 1 - lines of constant electrical field strength over the rectangular ϕ-Θ-plane (Θ-polarization)

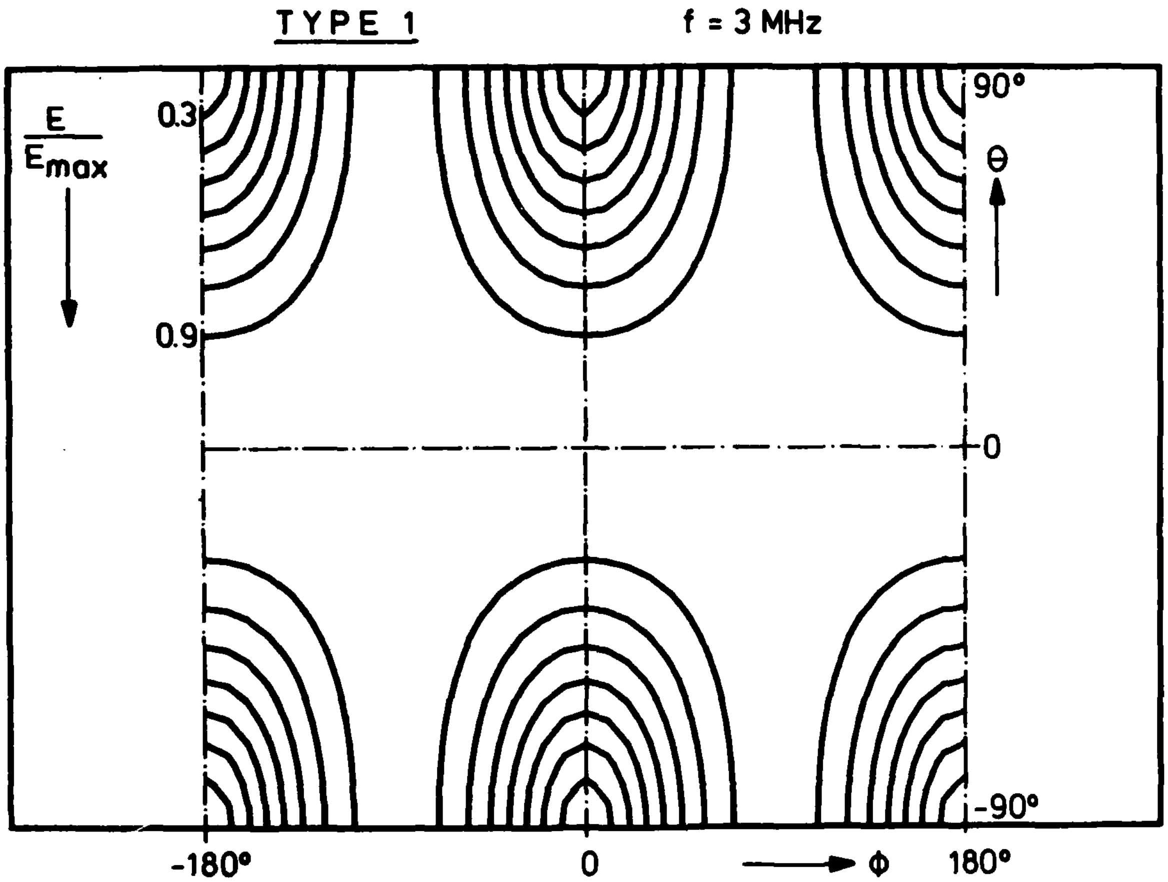


Fig. 23: Antenna type 1 - lines of constant total electrical field strength over the rectangular ϕ-Θ-plane

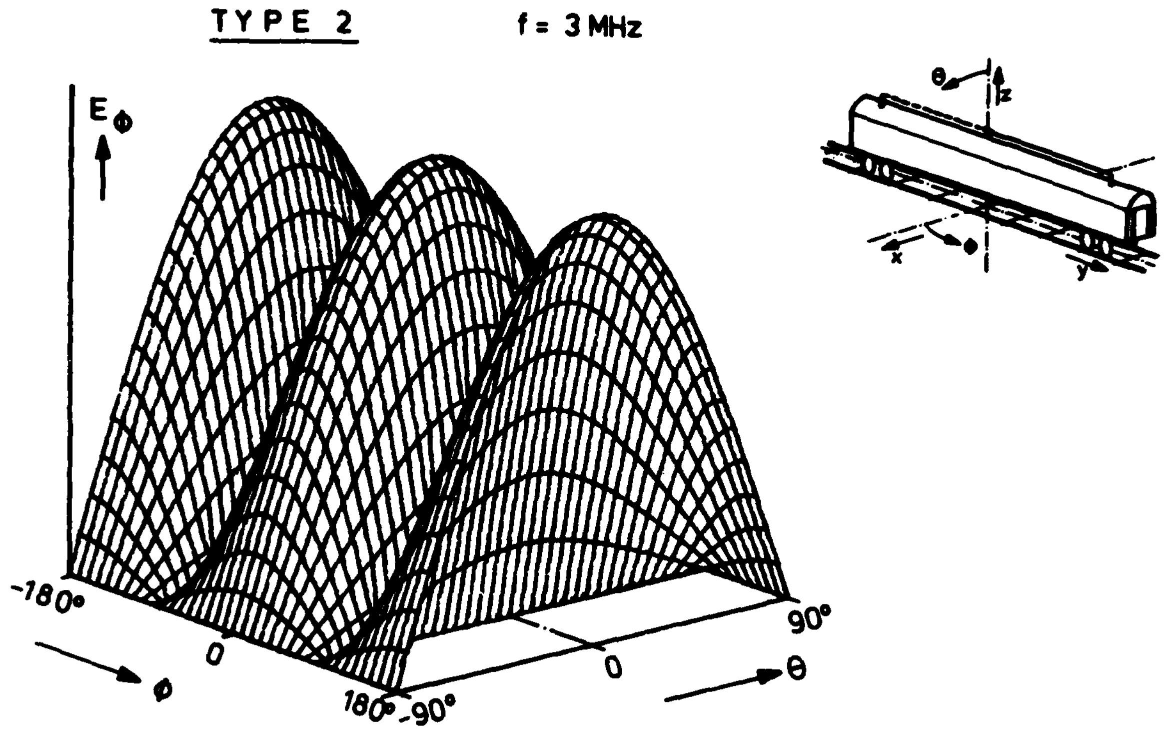


Fig. 24: ϕ-polarized electrical field strength

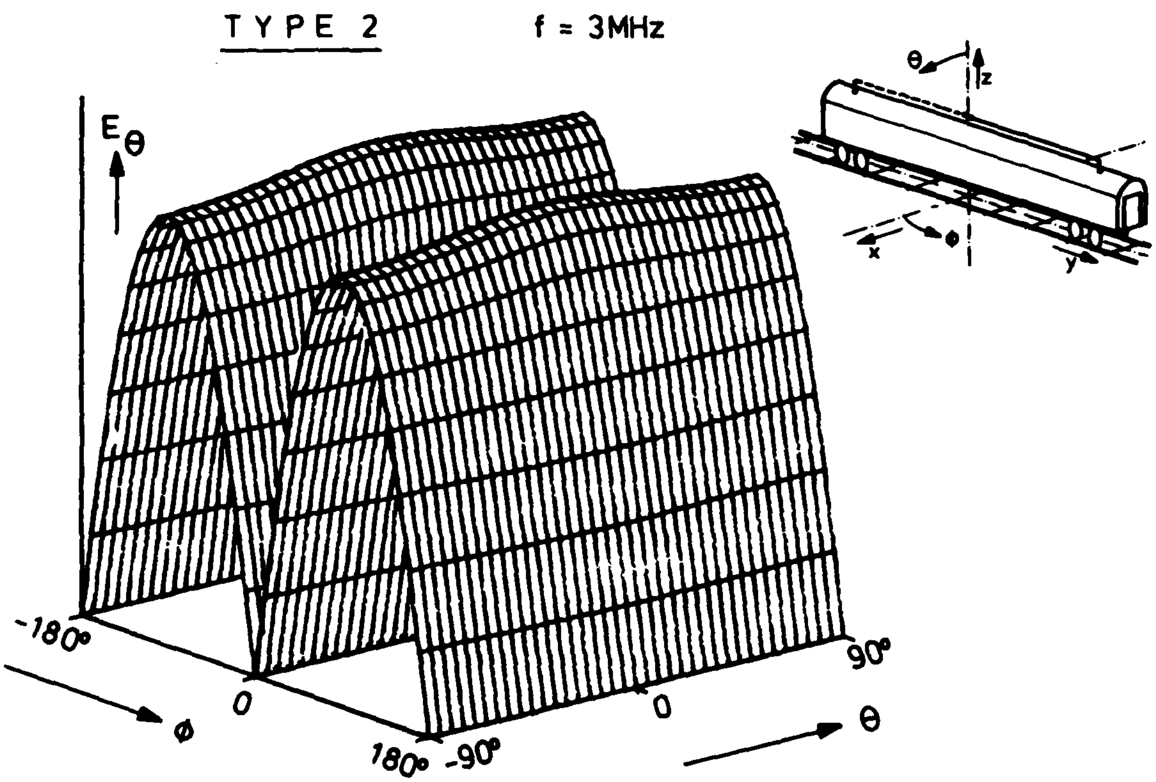


Fig. 25: Θ-polarized electrical field strength

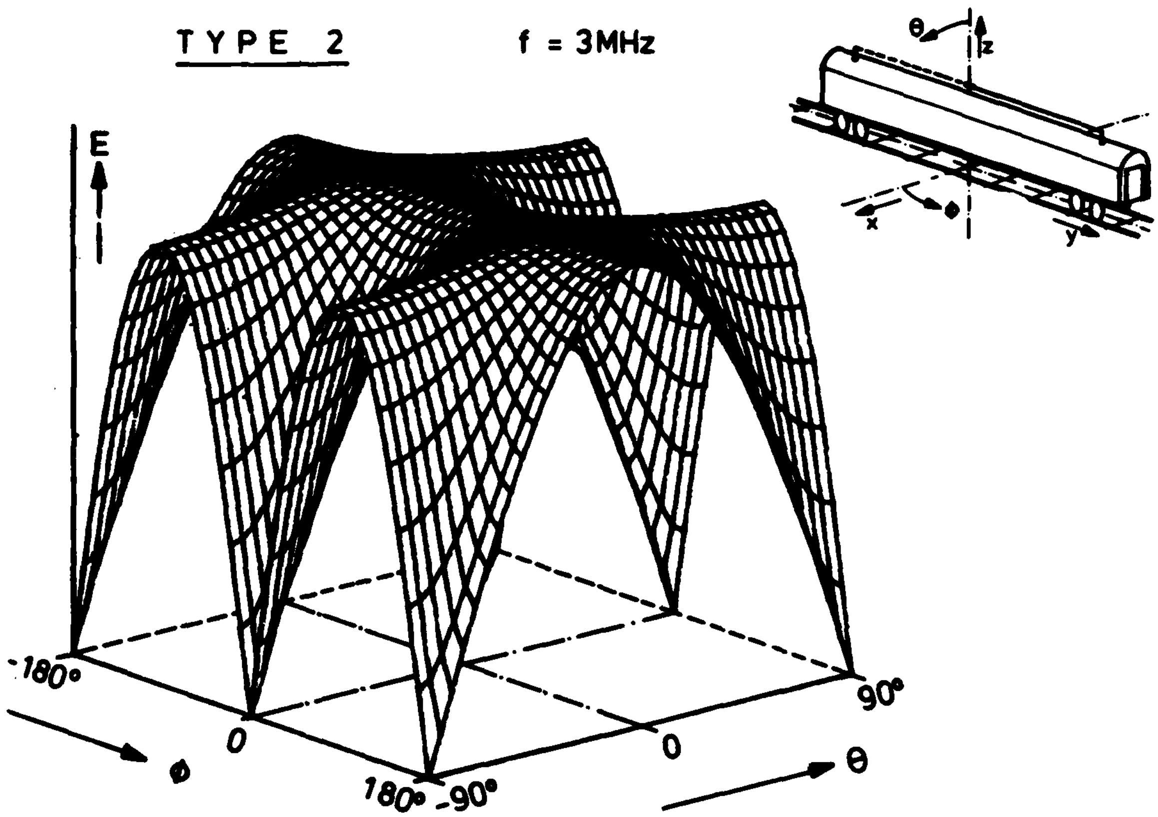


Fig. 26: Antenna type 2 - total electrical field strength over the rectangular ϕ-Θ-plane

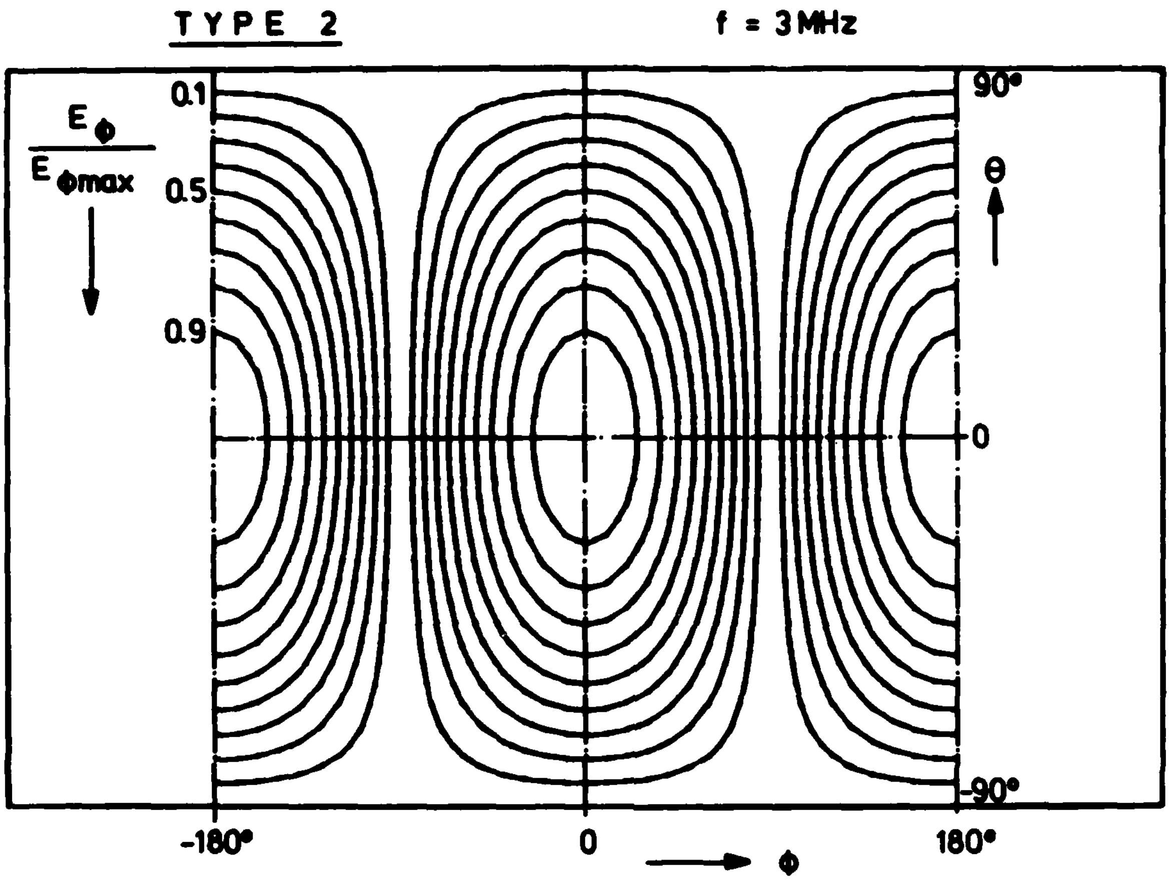


Fig. 27: Antenna type 2 - lines of constant electrical field strength over the rectangular φ-Θ-plane (φ-polarization)

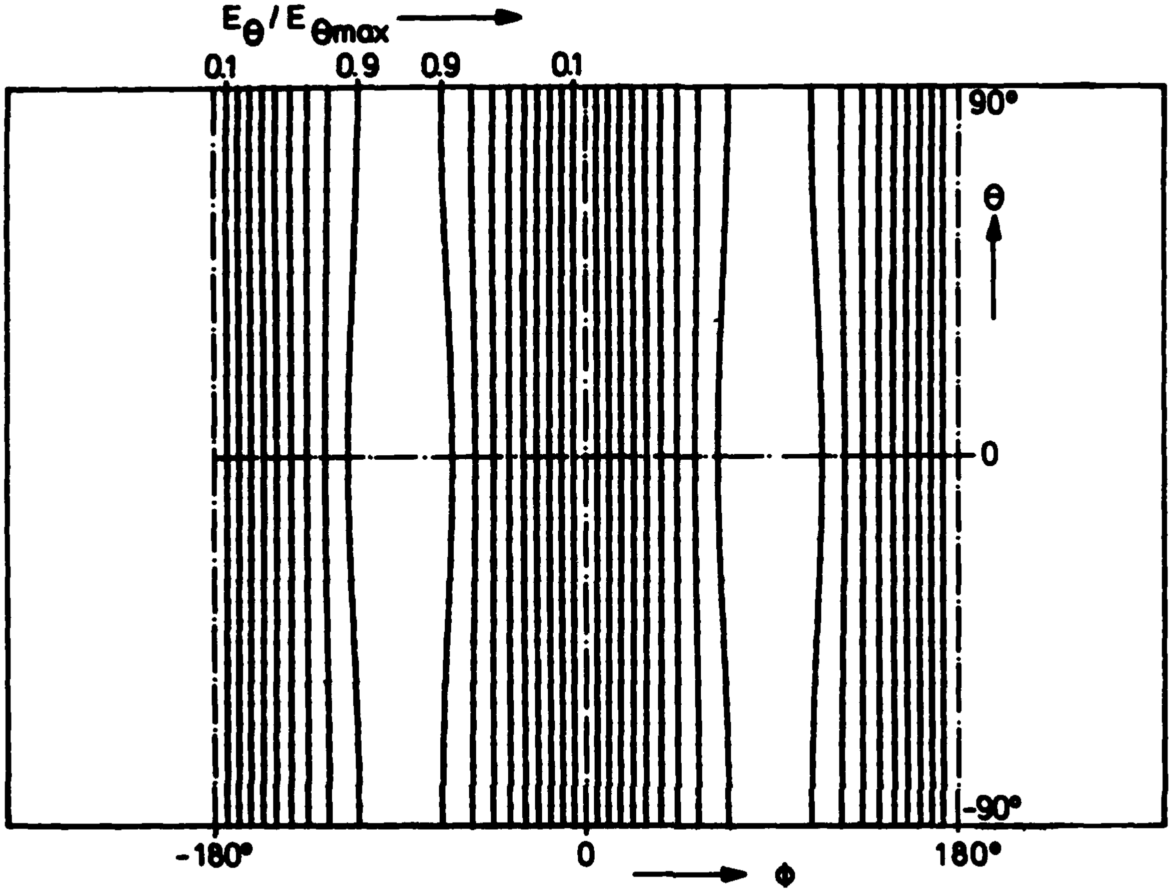


Fig. 28: Antenna type 2 - lines of constant electrical field strength over the rectangular ϕ-Θ-plane (Θ-polarization)

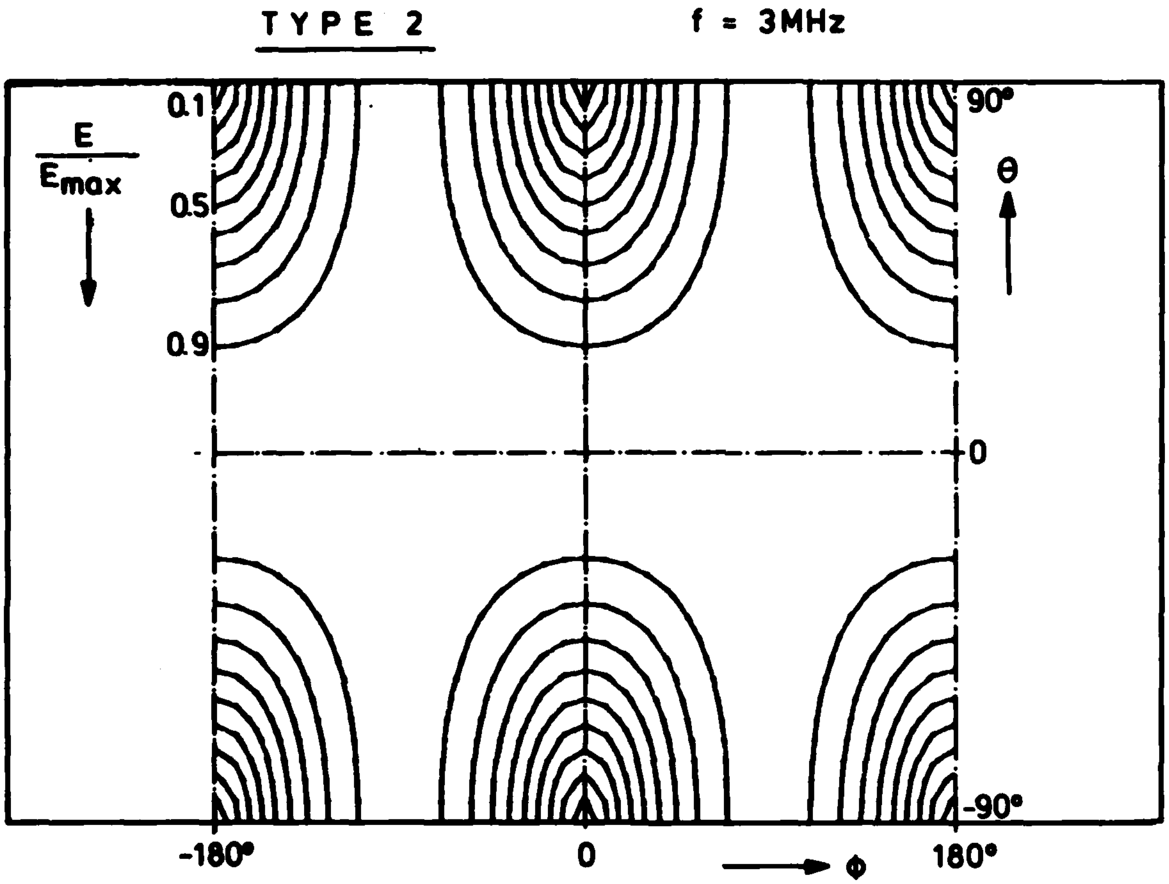


Fig. 29: Antenna type 2 - lines of constant total electrical field strength over the rectangular φ-Θ-plane

$$P_r = \frac{1}{2} \cdot |\underline{I}(o)|^2 \cdot R_{rs} \tag{19}$$

where R_{rs} is the unknown radiation resistance of the antenna, or in other words, the real part of the antenna impedance without losses. To make the integral of eq.17 applicable for a digital computer it is changed to a sum. In this case the radiation resistance can be written as:

$$R_{rs} = \frac{R_o^2}{|\underline{I}(o)|^2 \cdot Z_i} \cdot \sum_{\phi_j=0}^{\phi_j=2\pi} \cdot \sum_{\theta_i=0}^{\theta_i=\pi} (E^2 \cdot \sin\theta_i \Delta\theta) \cdot \Delta\phi \tag{20}$$

where: $\phi_j = \phi_{j-1} + \Delta\phi$

and: $\theta_i = \theta_{i-1} + \Delta\theta$

This radiation resistance would occur if a symmetric antenna configuration as shown in Fig.4 would be used. As the lower part of the antenna is only the image of the real antenna, the radiation resistance R of this real antenna, radiating only into the upper half-sphere is half the resistance of the symmetric configuration:

$$R_r = \frac{1}{2} \cdot R_{rs} \tag{21}$$

Although the boundaries in eq.20 can become smaller because of the symmetry of $E(\phi,\theta)$, the computer takes a lot of time to calculate R_r. If the whole antenna circuit shall be optimized, the calculation of R_r will increase the optimization procedure in an unjustifiable manner. Because of this - beside the numerical integration of eq.20 - an approximation for evaluating R_r has been used:

Antenna type 1 radiates in the total half-sphere, nearly like a half-wavelength dipole with its axis in the x-direction. The field strength of the far-field is given by eq.11 and eq.9, respectively, and is assumed to be approximately zero in those directions, where actually the minimum radiation occurs. This assumption seems to be quite rough, as for example at 4MHz the field strength for $\phi = 0^\circ$ and $\phi = 180^\circ$ is still about half the maximum field strength of $\phi = \pm 90^\circ$. But as the radiation resistance is in the milliohm range, this error will be negligible against the errors of estimating contact resistances or surface roughness, e.g.

The maximum electric field strength in the far-field of antenna type 1 is:

$$E_{max} = |\underline{I}(o)| \cdot Z_i \frac{h}{R_o \cdot \lambda_o} \cdot \tan(\beta_o \cdot a) \tag{22}$$

With the asumption of no radiation in the directions $\phi = 0^o$ and $\phi = +180^o$ the integration of the average Poynting vector over the radiation sphere leads to:

$$R_r \approx \frac{4}{3} \cdot \pi \cdot Z_i \cdot \left[\frac{h \cdot \tan(\beta_o \cdot a)}{\lambda_o}\right]^2 \tag{22a}$$

This equation can be modified to a more convenient formula:

$$R_r \approx 17.5 m\Omega \cdot \left[\frac{h}{[m]} \cdot \frac{f}{[MHz]} \cdot \tan(\beta_o \cdot a)\right]^2 \tag{22b}$$

At lower frequencies this result fits quite well with the numerical solution of eq.20 (see Fig.30 and Fig.31).

The radiation pattern of antenna type 2 is almost identical to that of a half-wavelength dipole with its axis in the x-direction. The maximum field strength in the far-field region can be found, using eq.13 and eq.9:

$$E_{max} = |\underline{I}(o)| \cdot Z_i \cdot \frac{2 \cdot h}{R_o \cdot \lambda_o} \cdot \tan(\beta_o \cdot a) \tag{23}$$

The integration of the average Poynting vector over the radiation sphere leads to:

$$R_r \approx \frac{16}{3} \cdot \pi \cdot Z_i \left[\frac{h \cdot \tan(\beta_o \cdot a)}{\lambda_o}\right]^2 \tag{24}$$

and can also be modified to a more convenient equation:

$$R_r \approx 70 m\Omega \cdot \left[\frac{h}{[m]} \cdot \frac{f}{[MHz]} \cdot \tan(\beta_o \cdot a)\right]^2 \tag{24a}$$

The dependence of R_r from height and length of the antenna and from the frequency is shown in Fig.32 and Fig.33. The results correspond well with the numerical solution of eq.20.

For both antennas, the dependence of the radiation resistance on height, frequency and length is described by a square law. The approximations which have been used limit the application of the formulas to a length which is smaller than a quarter wavelength.

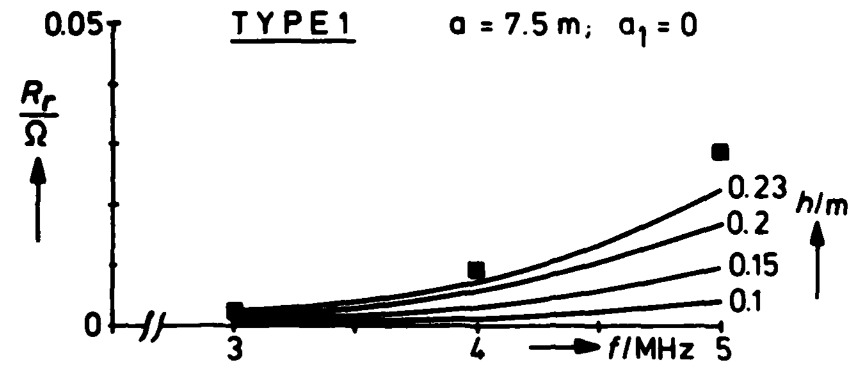


Fig. 30: Calculated radiation resistance of antenna type 1, depending on the frequency f and the height h of the current filament I.

■: result of numerical integration (h = 0.23)
Solid lines: approximation (eq. 22)

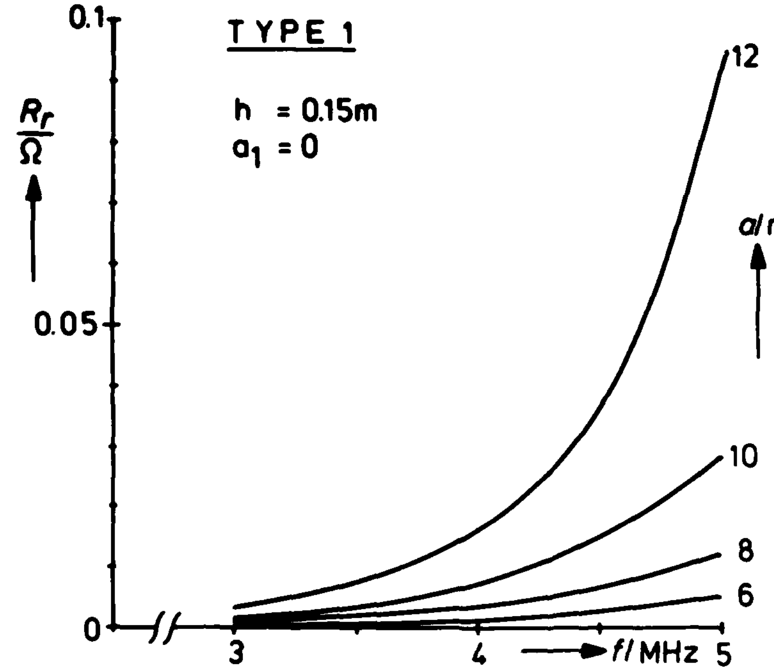


Fig. 31: Calculated radiation resistance of antenna type 1, depending on the frequency f and the length a of the antenna elements (approximation)

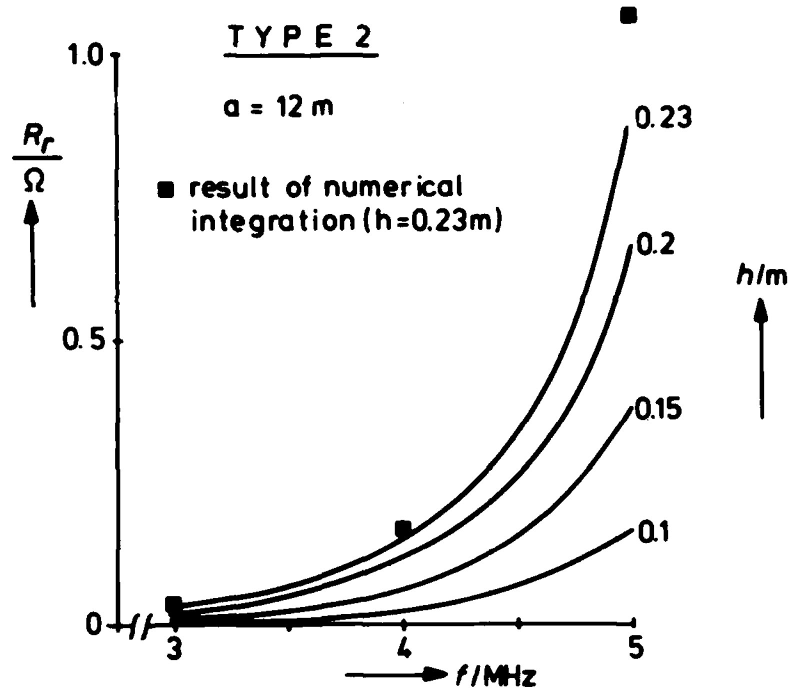


Fig. 32: Calculated radiation resistance of antenna type 2, depending on the frequency f and the height h of the current filament I (approximation)

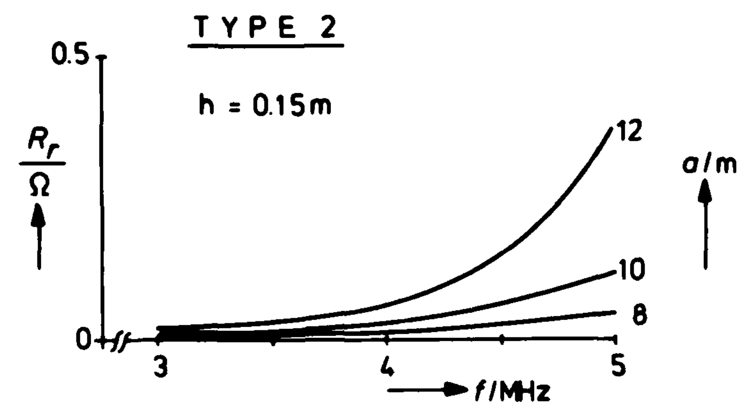


Fig. 33: Calculated radiation resistance of antenna type 2, depending on the frequency f and the length a of the antenna elements (approximation)

Since eq.22 is very similar to eq.24, both resistances can be compared if height and length are identical. Calling the resistance of antenna type 1 "R_{r1}" and that of antenna type 2 "R_{r2}", the quotient of both is

$$\frac{R_{r2}}{R_{r1}} \approx 4 \tag{25}$$

This result is quite reasonable as the area of antenna type 2 is two times the area of antenna type 1 and as the radiation resistance of electrical small loop antennas depends on the square of the area. With fixed element length of antenna type 1 and variable length a of antenna type 2 the ratio of the radiation resistances of the two antennas is shown in Fig.34.

2.3 Imaginary part of the antenna impedance

If the length a of the antenna elements is smaller than a quarter wavelength, the radiation resistance will be significantly smaller than the imaginary part of the antenna impedance. The imaginary part of the antenna impedance therefore is approximately that of a shortened lossless line. For antenna type 1 one obtains:

$$\begin{aligned} \underline{Z}_{A1} &= R_{r1} + j \cdot X_{A1} \\ &\approx R_{r1} + j \cdot Z_o \cdot \tan(\beta_o \cdot a) \end{aligned} \tag{26}$$

Antenna type 2 consists of two identical lines in series. Therefore the impedance of this antenna becomes:

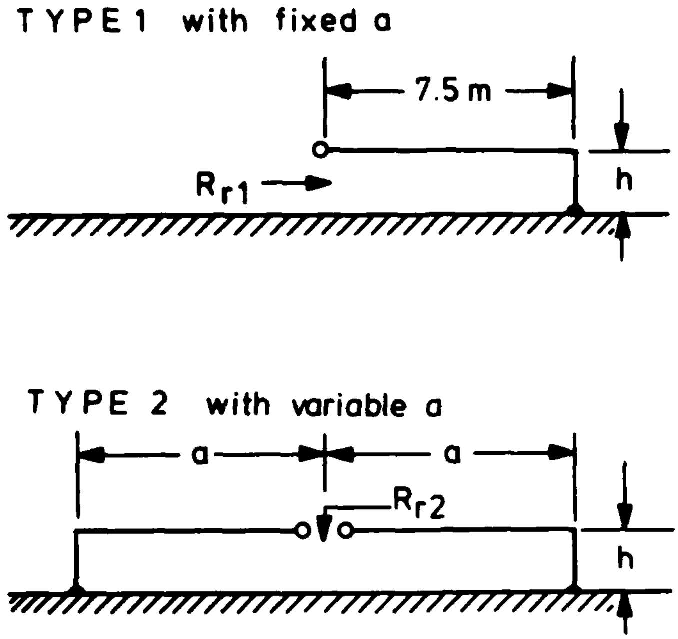
$$\begin{aligned} \underline{Z}_{A2} &= R_{r2} + j \cdot X_{A2} \\ &\approx R_{r2} + j \cdot 2 \cdot Z_o \cdot \tan(\beta_o \cdot a) \end{aligned}$$

The characteristic impedance Z_o of the line depends on the diameter d of the antenna elements. For a single element it is:

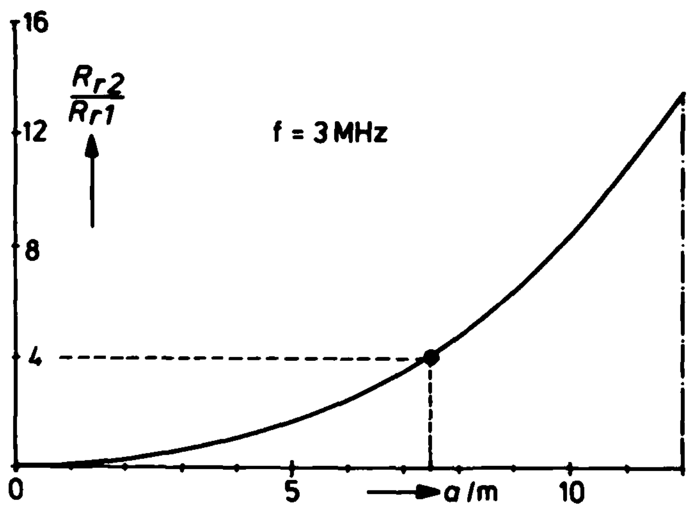
$$Z_{os} = 60\Omega \cdot \ln[2 \cdot \frac{h + h_{max}}{d} - 1]$$

Each tested antenna consisted of three elements in parallel. The resulting characteristic impedance of n conductors in parallel is usually not Z_{os}/n, as the fields of the parallel lines are not independent. For evaluating the characteristic impedance of the antenna elements in parallel a graphic method and an antenna model at higher frequencies has been used.

The ratio of the three-element line impedance Z_o and the single



a) Antenna configurations



b) Relation of the radiation resistances versus element length of antenna type 2

Fig. 34: Comparison of the radiation resistances of antenna type 1 and type 2, if length a of antenna type 1 is fixed

element characteristic impedance Z_{os} depends approximately only on the distance b of the elements if the outlines of the antenna are not changed (see Fig.35). The ratio has to be unique, if b/h approaches zero, as the line then becomes a single element line. If b is much greater then the height h of the current filament, the fields of the lines will become independent and the ratio Z_o/Z_{os} will approach $Z_o/Z_{os} = 1/3$. A sufficient approximation has been found by use of the equation:

$$\frac{Z_o}{Z_{os}} \approx \frac{1}{3} \cdot [1 + 2 \cdot \exp(- 0.7 \cdot \frac{b}{h})] \qquad (29)$$

This approximation will fail, if the feeding point is chosen close to one antenna element while the other two elements are connected via a longer conductor, representing a serial inductance which cannot be neglected at higher frequencies.

The far-field of the three-element antenna does not differ from that of the single element antenna, as the currents in the antenna elements have the same direction each and the path difference Δ is small, compared to the wavelength λ_o.

In the realized version, antenna type 1 consists not only of a shortened line but also of an open-ended part of the length $a_1 = 1.5$m (see Fig.1). As this open line is shorter than a quarter wavelength it acts as a capacitor in parallel to the feeding point. The current into this line is very small compared to the current into the shortened line and can be neglected for the calculation of the far-field. For the evaluation of the antenna's imaginary part, however, it has to be taken into account. Therefore, the imaginary part of the antenna impedance of type 1 becomes:

$$X_{A1} = Z_o \cdot \frac{\tan(\beta_o \cdot a)}{1 - \tan(\beta_o \cdot a) \cdot \tan(\beta_o \cdot a_1)} \qquad (30)$$

2.4 Losses

The losses of the antenna are caused by:

- the contact resistances between coupler and antenna

- the serial resistance of the antenna element, which is determined by inhomogenious current distribution, skin effect, and roughness of the conductor surface. All of these parameters are increased by corrosion

- the contact resistances of the grounding strips

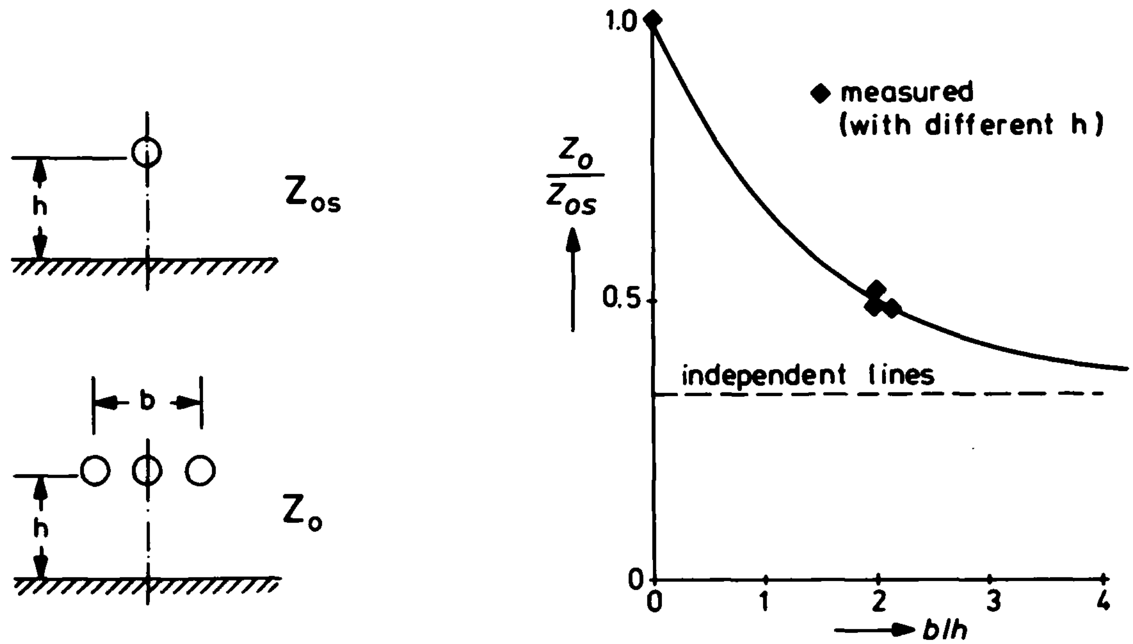


Fig. 35: Characteristic impedance Z_o of a 3-element line, compared to the impedance Z_{os} of a single element line

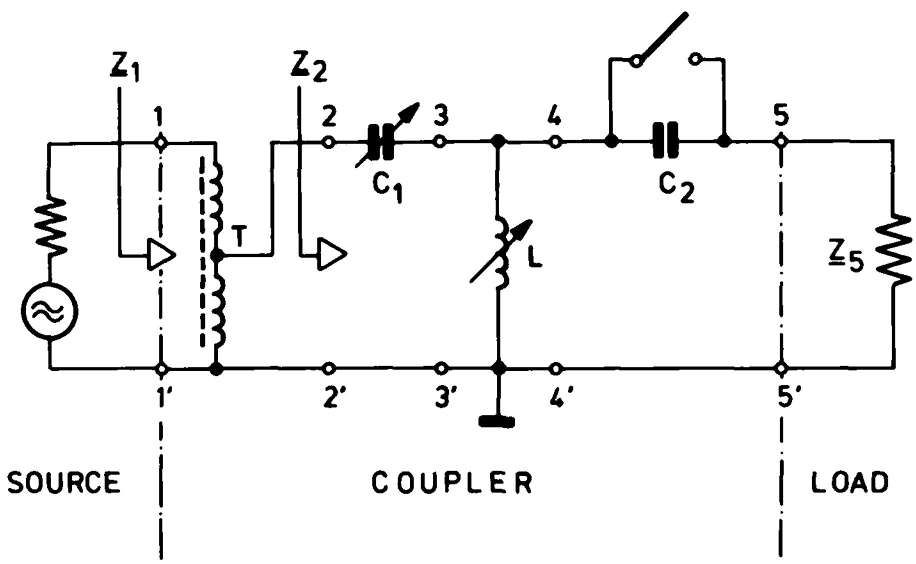


Fig. 36: Rf-circuit diagram of the HARRIS-antenna coupler RF615B

- the parallel resistances of the insulators between antenna and roof, when covered with coal particles, dust or dirty water

- the parallel resistance over the total antenna length, when snow and ice, mixed with dirt, covers the roof.

Most of the presented losses can only be estimated and will change with aging, weather conditions, corrosion, and maintenance. For the calculation of the antenna efficiency the best values that can be obtained have been used.

The connection between coupler and antenna has to be made of a material with low losses and a low propensity for corrosion. The contacts, coupler-connector and connector-antenna, should be protected against wetness to avoid electro-chemical corrosion. Under these conditions a resistance of about $R_c \approx 20m\Omega$ for each contact seems to be reasonable. The power loss caused by these resistances depends on the input current I(0) of the antenna and is for each contact:

$$P_{L1} = \frac{1}{2} \cdot |\underline{I}(o)|^2 \cdot R_c \qquad (31)$$

The loss caused by the finite conductivity of the antenna elements can only be found by integrating the differential losses over the total antenna length, as the current density is a function of y (see eq.2). The total loss of one antenna element can be substituted by a serial resistance, located at the antenna's feeding point. As there are three elements connected in parallel, the total resistance R is one third of the resistance of a single element and becomes:

$$\begin{aligned} R_a &= \frac{1}{3} \cdot \frac{R'}{\cos^2(\beta_o \cdot a)} \cdot \int_{y=o}^{y=a} \cos^2[\beta_o \cdot (a-y)] \cdot dy \\ &= \frac{1}{6} \cdot R' \cdot a \cdot \frac{2 \cdot \beta_o \cdot a + \sin(2 \cdot \beta_o \cdot a)}{2 \cdot \beta_o \cdot a + \cos^2(2 \cdot \beta_o \cdot a)} \end{aligned} \qquad (32)$$

where R' is the surface resistance of the antenna element

$$R' = \frac{2 \cdot h_{max} - d}{2 \cdot h \cdot d} \cdot \sqrt{\frac{f \cdot \mu}{\sigma \cdot \pi}} \qquad (33)$$

(μ = permittivity of antenna element and roof
σ = conductivity of antenna element and roof)

If the surface of the conductors is rugged, the resistance will grow with the factor F_R:

$$R'' = R' \cdot F_R$$

$$F_R = 1 + \frac{2}{\pi} \cdot \arctan[1.4 \cdot (\frac{\Delta r}{\delta})^2] \tag{34}$$

where Δr is the RMS-value of the surface ruggedness and δ is the penetration depth:

$$\delta = \frac{1}{\sqrt{\pi \cdot f \cdot \sigma \cdot \mu}} \tag{35}$$

The imaginary part of the surface resistance can be neglected against the imaginary part of the short circuited line.

The dissipative power, caused by the finite conductivity of the antenna element is:

$$P_{La} = \frac{1}{2} \cdot |\underline{I}(o)|^2 \cdot R_a$$

The contact resistance of the grounding strips strongly influences the total efficiency since the antenna current has a maximum at the short. The strips have to be as short as possible and should be soldered and protected against wetness and corrosion. Only if the short is connected very carefully in the way described before, can the contact losses be neglected at this point.

If the insulators which bear the antenna elements at the roof are clean, no parallel losses will occur. If the insulators are covered with some conducting deposits, for example from the exhaust fumes of the engine, the losses will depend on the position of the insulator, the length of the insulating gap, the cross-section of the conducting cover, and on its conductivity. To avoid this kind of loss the insulator gap has to be as broad as possible, protected by a shield and cleaned at certain time intervals.

As the antenna height is very small, snow and ice mixed with conducting particles may decrease the efficiency of the antenna. In these cases the roof has to be cleaned.

2.5 Problems

With the results of the antenna calculation, it is possible to estimate the voltages and currents, which occur at the antenna. For a radiated power of only 2 Watts with antenna type 1, a current with the magnitude of $I(0) = 17{,}5A$ and a voltage of $U(0) = 1.29kV$ is necessary, as the phase angle between voltage and current is close to 90° !

Because of the high currents, nonlinear effects such as the generation of harmonics as well as intermodulation may occur at the connections of the antenna elements.

3. Transmission Line

When the antenna has been successfully tuned, the transmission line between coupler and transmitter will be connected to a matched load. Because of this and the small length of the line, the losses are small too.

Antenna type 2 uses a second transmission line between the coupler and the symmetric feeding point which actually transforms the antenna impedance. Since the current at the feeding point has to be very high, the serial cable losses will not be negligible. The following calculations apply to antenna type 2 only.

3.1 Losses of the coaxial feeder line.

The coaxial feeder line should be filled with air and have only a few dielectric supports. On this account, the losses caused by the electric field are negligible. The current density is nonhomogenous as the magnitude of the antenna's reflection coefficient is very close to 1. The total loss of the coaxial line can be evaluated by integrating the differential serial losses over the full length of the line. The total loss can be substituted by a serial resistance R_{LC}, which is very small compared to the imaginary part of the antenna impedance. This fictitious loss resistance is located at the symmetrical feeding point:

$$R_{LC} = \frac{R_c'}{\cos^2(2\cdot\pi\cdot\frac{l_o}{\lambda_\varepsilon})} \cdot \int_{z=l_o}^{z=l_o+l_T} \cos^2(2\cdot\pi\cdot\frac{z}{\lambda_\varepsilon})\cdot dz \qquad (37)$$

where: R'_c = surface resistance of the coaxial line:

$$R_c' = (\frac{1}{D_a} + \frac{1}{D_i}) \cdot \sqrt{\frac{f\cdot\mu_o}{\pi\cdot\sigma}}$$

D_a = inner diameter of outer conductor
D_i = outer diameter of inner conductor
l_o = length of a hypothetical coaxial line, short circuited and with the same characteristic impedance Z_o as the feeder line and the same input impedance as the antenna (real part neglected)

$$l_o = \frac{\lambda_\varepsilon}{2\cdot\pi}\cdot \arctan(X_{A2}/Z_o)$$
$$Z_o = 60\Omega\cdot \ln(D_a/D_i) \tag{39}$$

l_T = mechanical length of coaxial line = length a of antenna element

The wavelength λ_ε at the coaxial line is very close to the free-space wavelength, as the dielectric supports do not appreciably change the phase velocity.

The solution of the integral leads to:

$$R_{LC} = R'_c\cdot\lambda_\varepsilon\cdot\frac{4\pi\cdot\frac{l_T}{\lambda_\varepsilon} - \sin(4\cdot\pi\cdot\frac{l_o}{\lambda_\varepsilon}) + \sin(4\pi\cdot\frac{l_o + l_T}{\lambda_\varepsilon})}{8\cdot\pi\cdot\cos^2(2\cdot\pi\cdot\frac{l_o}{\lambda_\varepsilon})} \tag{41}$$

3.2 Transformation of the coaxial line

The radiation resistance and the loss resistance of the antenna as well as the contact resistances and the loss resistance of the coaxial line had been transformed to serial resistances located at the balanced feeding point of the antenna. The summed-up resistances are still much smaller then the imaginary part X_{A2} of the antenna impedance:

$$\underline{Z}_6 = (R_{r2} + R_a + R_c + R_{LC}) + jX_{A2} \tag{42}$$

The coaxial line which now theoretically is assumed to be lossless, as its losses are represented by the resistance R_{LC}, transforms the impedance $\underline{Z}_6$ to the impedance $\underline{Z}_5$:

$$\underline{Z}_5 = \frac{\underline{Z}_6 + jZ_o\cdot\tan(2\pi\cdot\frac{l_T}{\lambda_\varepsilon})}{1 + j\frac{\underline{Z}_6}{\underline{Z}_o}\cdot\tan(2\cdot\pi\cdot\frac{l_T}{\lambda_\varepsilon})} \tag{43}$$

4. Antenna Coupler (HARRIS RF615B)

4.1 Circuit diagram and functions

The RF-circuit diagram of the antenna coupler is shown in Fig.36. The impedance Z_5 has to be transformed to an impedance of $\underline{Z}_1 = 50\Omega$. The antenna matching will be accepted by the coupler, as soon as the standing wave ratio at pins 1-1' is not greater than SWR = 2. Otherwise, the transmitter automatically will be connected to a dummy-load and an "overload" will be indicated. Capacitor C_1 and inductance L are motor controlled. Capacitor C_2 can be short-circuited by a relay. The tuning is done automatically by help of two discriminators in less than five seconds.

The transformer T has a voltage ratio of 2 and transforms the source-impedance of 50 Ohms at pins 1-1' to a source-impedance of 12.5 Ohms at pins 2-2'. The matching condition is:

$$\underline{Z}_2 = 12.5\Omega$$

The losses of the toroid are usually negligible.

4.2 Losses of the coupler elements

Capacitor C_1 is a vacuum type which can be varied between 12pF and 500pF. The losses caused by the electric field are negligible, but there are serial losses caused by the current, especially at the rotor contact. These losses have been substituted with a 20-Milliohm serial resistance (see Fig.37).

The variable coil is a cylindrical inductor, the unused part of which is short-circuited. The tuning range of the inductance L has been assumed to be $1\mu H \leq L \leq 50\mu H$. The Q-factor Q of the coil depends on the serial resistance of the rotor contact, the losses of the magnetic field, the losses caused by the current, including skin effect and proximity effect and losses from the coupling of the shortened part of the coil and the part in use. This coupling effect usually shows maximum values at some resonant frequencies.

As known from several different couplers the Q-factor varies with the tuning position and the frequency. The dependence of the Q-factor from the rotor position and the frequency was assumed to be:

$$Q_L = Q_o \cdot \sqrt{\frac{L}{L_{max}} \cdot \frac{f}{f_{max}}} \qquad (44)$$

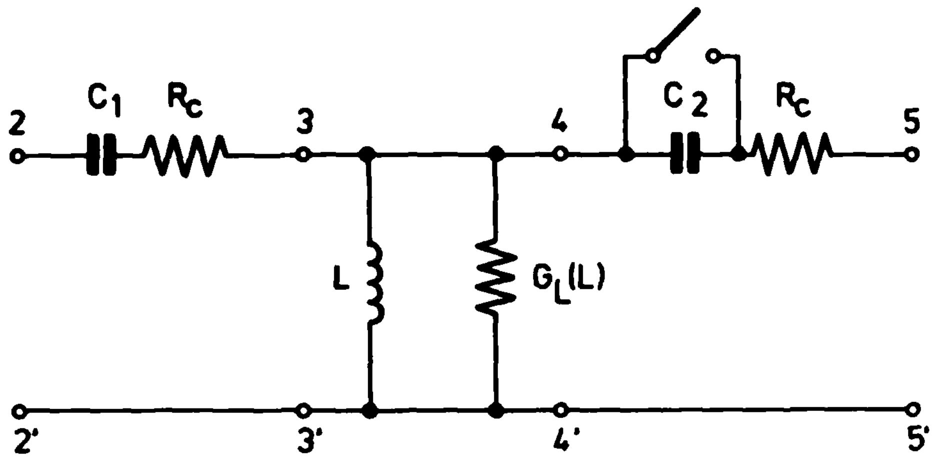


Fig. 37: Circuit diagram of the coupler between Pins 2-2' and 5-5' including loss resistances

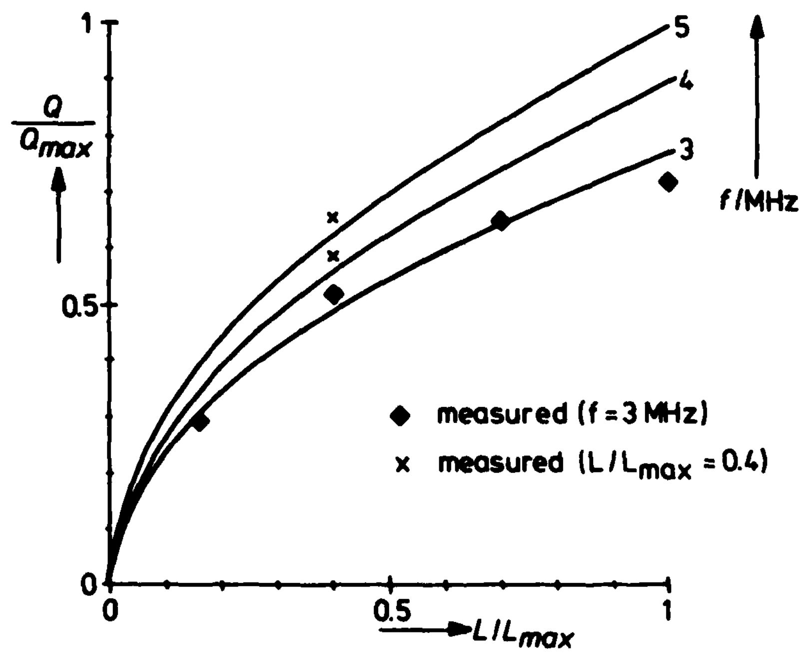


Fig. 38: Q-factor of the variable inductor L versus inductance. Parameter: frequency f. solid lines: approximation (eq. 44)

where Q_o was estimated to be $Q_o = 380$ with $L_{max} = 50\mu H$ and $f_{max} = 5MHz$. Eq.44 is quite reasonable, as the length of the coiled wire increases with the number w of turns, while the inductivity increases with w^2. Furthermore, the reactive power increases linearly with the frequency, while the losses, caused by the skin effect, increase with $\sqrt{f}$:

$$\left.\begin{array}{lll} P_{reactive} & \sim & f \cdot L \\ P_{loss} & \sim & w \cdot \sqrt{f} \\ w & \sim & \sqrt{L} \\ Q_L = P_{reaktive}/P_{loss} & & \end{array}\right\} \quad Q_L \sim \sqrt{L \cdot f}$$

Fig.38 shows the dependance of the inductor's Q-factor from the frequency and the inductivity. The solid lines are the result of eq.44. The measurement results documented in this diagram have been achieved with a variable coil used in one of the investigated couplers.

The conductance G_L, representing the losses of the coil, can be calculated using eq.45:

$$G_L = \frac{1}{\omega \cdot L \cdot Q_L} \qquad (45)$$

The capacitor C_2 is a fixed ceramic capacitor (200pF) which can be by-passed by a relay contact. The losses of this capacitor and the contact resistance have been substituted by a serial resistor of R_c = 20 Milliohms.

The values estimated for the coupler's elements may vary widely, because of a damaged rotor or relay contact. The reliability of the estimations of the coupler losses can only be verified by comparision of the calculated and measured total efficiency of the antenna configuration, including the coupler.

4.3 Calculation of the coupler transformation

The computer program which has been used works similar to the coupler function. With the relay contact in parallel to C_2, open and closed respectively, the inductivity L is varied in small steps. For each value of L, the variable capacitor C_1 is tuned, if possible, to a value, where the imaginary part of $\underline{Z}_2$ is as close as possible to zero. The impedance $\underline{Z}_2$ is then transformed

to the input impedance $\underline{Z}_1$. With this impedance the standing wave ratio is calculated. If SWR ≤ 2, the currents, voltages, and powers of the antenna, the coaxial line, and the coupler is computed.

(The matching procedure can also be done by solving a 6-th order polynomial. This method has been tested too, but the step-by-step method described before can be supervised much easier and doesn't take much time. So we preferred this way).

The computation will now be described in more detail:

The impedance $\underline{Z}_4$, is, depending on the relay contact

$$\underline{Z}_4 = \underline{Z}_5 + R_c + \begin{cases} 0 & \text{(contact closed)} \\ \frac{1}{j\omega C_2} & \text{(contact open)} \end{cases} \tag{46}$$

where $\underline{Z}_5$ is known by eq.43.

The impedance $\underline{Z}_4$ is transformed into the admittance $\underline{Y}_4$:

$$\underline{Y}_4 = \frac{1}{\underline{Z}_4} \tag{47}$$

The admittance $\underline{Y}_4$ has to be added to the admittance of the lossy inductivity L:

$$\underline{Y}_3 = \underline{Y}_4 + G_L + \frac{1}{j\omega L} \tag{48}$$

and is then transformed into the corresponding impedance $\underline{Z}_3$. The capacitor C_1 has to compensate for the imaginary part X_3 of the impedance $\underline{Z}_3$:

$$C_1 \stackrel{!}{=} \frac{1}{\omega \cdot X_3} \tag{49}$$

As the variation range of the capacitor C_1 is between C_{min} = 12pF and C_{max} = 500pF the compensation may fail. In this case the most favorable value of C_1 is chosen (C_{min} or C_{max}). The impedance $\underline{Z}_2$ becomes:

$$\underline{Z}_2 = \underline{Z}_3 + R_c + \frac{1}{j\omega C_1} \tag{50}$$

The standing-wave ratio at pins 1-1' can be found by

transforming $\underline{Z}_2$ to $\underline{Z}_1$ with the transformer T and referring $\underline{Z}_1$ to R_Q:

$$SWR = \frac{1 + |\underline{r}|}{1 - |\underline{r}|} \tag{51}$$

where:
$$\underline{r} = \frac{4 \cdot \underline{Z}_2 - R_Q}{4 \cdot \underline{Z}_2 + R_Q} \tag{52}$$

If the SWR is less then SWR = 400, the computer stores this value. Otherwise, the probability of getting a SWR of less or equal SWR = 2 by a fine-tuning of L is extremely low. Among the stored values, the computer takes that one with the smallest SWR and optimizes it by a fine-tuning of L and C_1. The optimization stops at SWR < 1.002, or after 20 tuning routines.

5. Calculation of powers, voltages and currents

If the optimized SWR is not greater than SWR = 2 the currents, voltages and powers are computed for an available generator power of P_a = 100 Watts. The input power P_1 delivered to the coupler is

$$P_1 = P_a \cdot (1 - |\underline{r}|^2) \tag{53}$$

The amplitudes of the voltage $U_{i-i'}$ and the current I_i, referred to any pins i-i' are dependent on the power $P_{i-i'}$ delivered to these pins:

$$U_{i-i'} = \sqrt{\frac{2 \cdot P_{i-i'}}{G_{i-i'}}} \tag{54}$$

$$I_{i-i'} = \sqrt{\frac{2 \cdot P_{i-i'}}{R_{i-i'}}} \tag{55}$$

where: $G_{i-i'} = \mathrm{Re}\,\{\underline{Y}_{i-i'}\}$, and: $R_{i-i'} = \mathrm{Re}\,\{\underline{Z}_{i-i'}\}$

The power $P_{2-2'}$ at pins 2-2' is the same as P_1 since the transformer is assumed to be lossless. The current I_2 causes losses at the capacitor C_1, represented by the contact resistance R_c. On this account the power $P_{2-2'}$ flowing through pins 3-3' is:

$$P_{3-3'} = P_{2-2'} - \frac{1}{2} \cdot I_2^2 \cdot R_c \tag{56}$$

The current I_2 is found using eq.55.
The voltage $U_{3-3'}$ causes losses within the coil L represented by its parallel conductance G_L. The remaining power $P_{4-4'}$ is:

$$P_{4-4'} = P_{3-3'} - \frac{1}{2} \cdot U^2_{3-3'} \cdot G_L \qquad (57)$$

The power $P_{5-5'}$ is, because of the contact losses at C_2:

$$P_{5-5'} = P_{4-4'} - \frac{1}{2} \cdot I^2_4 \cdot R_c \qquad (58)$$

The power $P_{5-5'}$ feeds either directly (antenna type 1) or via the coaxial line (antenna type 2) to the antenna. Since the cable losses have been substituted by the serial resistance R_{Lc} at the antenna feeding point, the cable can be handled like a lossless line with the power $P_{6-6'}$ at the output:

$$P_{6-6'} = P_{5-5'}$$

The output current and voltage can be calculated using again eq.54 and eq.55. With antenna type 2, the losses of the line become:

$$P_{Lc} = \frac{1}{2} \cdot I^2_6 \cdot R_{Lc} \qquad (59)$$

as: $I_6 = I(0)$

The final radiated power is:

$$P_r = \frac{1}{2} \cdot I^2_6 \cdot R_r \qquad (59a)$$

The power lost by the surface-resistance of the antenna elements is:

$$P_{La} = \frac{1}{2} \cdot I^2_6 \cdot R_a$$

and the power lost by the contact resistances is:

$$P_c = \frac{1}{2} \cdot I^2_6 \cdot R_c \qquad (59c)$$

Since the available power is chosen to be 100 Watts, the calculated powers are identical to the percentage of total power available from the transmitter.

6. Optimization of the antenna

The parameters of the antenna which can be varied are:

- antenna type

- length a
- diameter d
- length a_1 of the parallel open ended line (type 1 only)
- characteristic impedance Z_o of the coaxial feeder line (type 2 only)

For each frequency (3, 4 and 5MHz) which has been investigated, a different optimized antenna could be found. The following figures show the parameter variation for the two antenna types.

Fig.39 represents the dependence of the total efficiency of antenna type 1, including the HARRIS-coupler, on the length of the antenna elements and the transmitter frequency f. The diagram shows:

1. The efficiency increases with the frequency, mainly because of the increasing Q-factor of the coil.

2. The longer the antenna, the better the efficiency.

The diagram documents that maximum efficiency will occur at a frequency of 5MHz if the length of the antenna elements is l = 15m. On the other hand, the coupler will not be able to match the antenna at a frequency close to resonance. Therefore the maximum length of the antenna is only 16m for f = 3MHz, 10m for f = 4MHz and 6m for f = 5MHz. If the whole frequency range from 3 MHz up to 5 MHz shall be used, the maximum length of the antenna has to be a = 6m to keep it tunable. With this length, a diameter of d = 7.5cm, and no parallel line, the efficiency will be about 0.4% at 3MHz, 1% at 4MHz and 2.7% at 5MHz.

Fig.40 shows the total efficiency of the antenna configuration, depending on the antenna diameter d. The efficiency decreases with increasing diameter. Therefore the lower boundary seems to be the mechanical stability of the antenna.

The length a_1 of the parallel open-ended line, up to a_1 = 4m has nearly no influence on the antenna efficiency, as shown in Fig.41. So these elements should be omitted.

If the three parallel antenna elements are spaced closer together, the antenna's characteristic impedance is increased and the efficiency can be raised further.

The optimum sized antenna of type 1 has to have:

- a length as long as possible for matching
- a diameter as small as possible for matching and as justifiable for stability

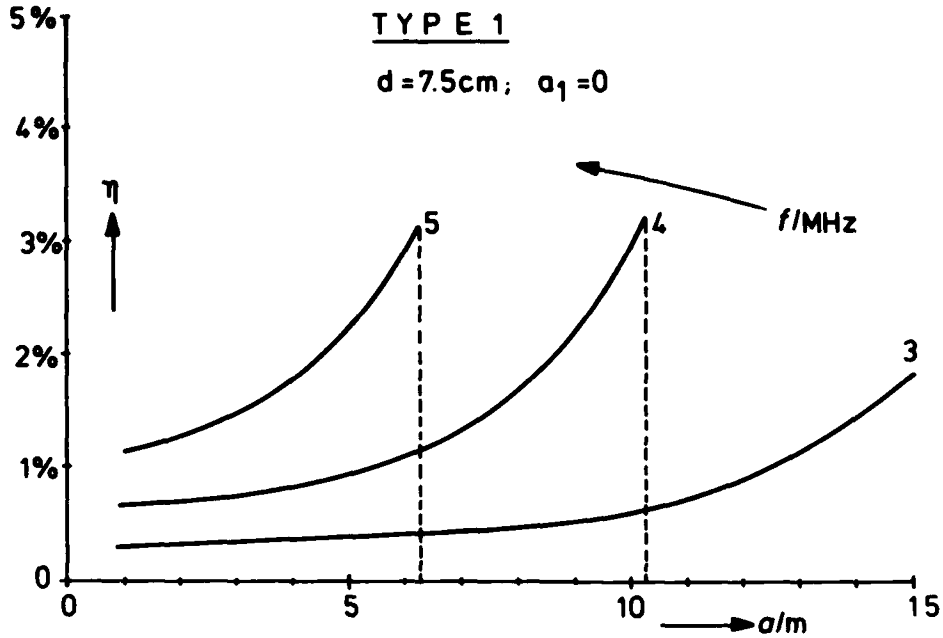


Fig. 39: Total efficiency η of antenna type 1 depending on the length a of the antenna elements and on the frequency f

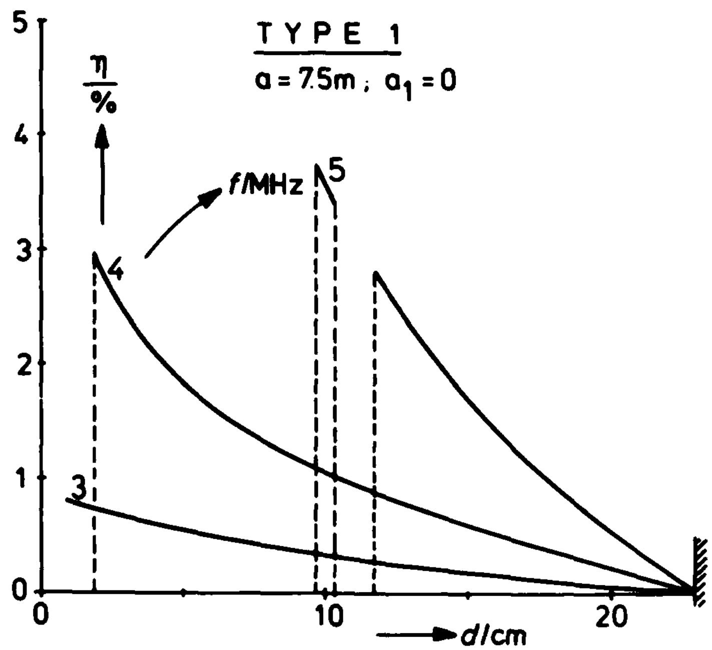


Fig. 40: Total efficiency η of antenna type 1, depending on the diameter d of the antenna elements and on the frequency f

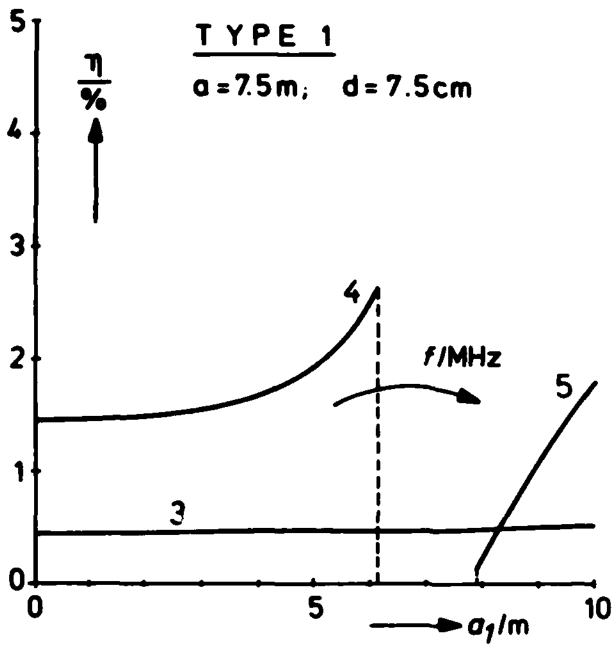


Fig. 41: Total efficiency of antenna type 1, versus length a_1 of the open-ended parallel line. Parameter: frequency f

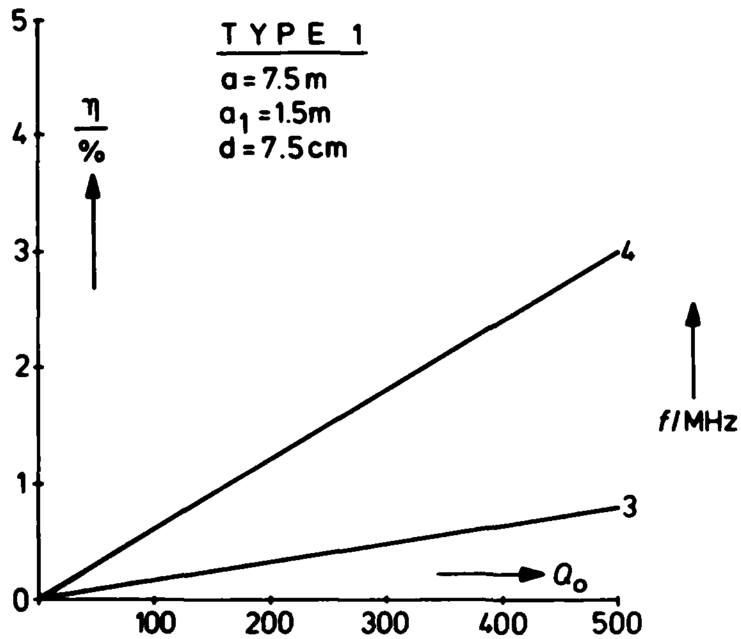


Fig. 42: Total efficiency η of antenna type 1 versus maximum Q-factor Q_o of the variable inductor L. Parameter: frequency f

- the antenna elements mounted very close together.

For example, an aluminum construction with a length of $a = 10m$, an open-ended parallel line of the length $a_1 = 0$, a diameter of $d = 4cm$ and a total width of $b = 40cm$ might fulfill these demands and would result in a total efficiency of 1% at $f = 3MHz$ or a gain of more then 4dB compared to the currently existing antenna. Unfortunately, this antenna cannot be matched at 4MHz and 5MHz with the applied coupler. That means that the optimized antenna for the critical, i.e. the lowest frequency, is not tunable at the remaining frequencies. This tendency can be seen with both antenna types. Therefore a compromise must be made between efficiency at the lowest frequency and tunability at the remaining frequencies. Antenna type 1 in the existing realization has an efficiency of about 0.4% at 3MHz and is an acceptable solution. But even with these dimensions the coupler will not be able to tune the antenna at the highest frequency.

Antenna type 2 allows a much higher efficiency than type 1. At 3MHz the efficiency can be calculated to be greater than 12 percent for an antenna with the element length $a = 10m$, the diameter of the elements of $d = 13cm$ and a 50-Ohm feeder line. If the diameter is increased a small amount, the antenna can no longer be tuned. Fig.46 documents the tuning problem. By increasing the length of the antenna elements, the current through the coupler's inductivity L is made smaller and smaller, while simultaneously its Q-factor becomes better. This improvement is only usable until the maximum inductance of the variable coil L is attained. Better efficiency could be possible by choosing a longer antenna, but the built-in inductivity cannot be tuned to the optimimum antenna type. Unfortunately, the maximum ranges and the non-tunable ranges at the different frequencies are overlapping. With the remaining parameters fixed, as stated in Fig.43, the antenna is only tunable at all three frequencies if the length a is greater than 10m. The achievable efficiency with these antenna dimensions will be between 1.6% to 5.5% - and so still much better then the currently existing antenna.

As described before, the efficiency of antenna type 1 is increased by decreasing the diameter of the antenna elements. As documented in Fig.44, the influence of the diameter on the efficiency of antenna type 2 is vice-versa than with type 1. At 3 MHz, a maximum occurs with an efficiency of about 8%, when $d = 13cm$. With a diameter between $13cm < d < 19cm$ or greater than $d = 21cm$, the antenna is not tunable. At $f = 4MHz$ the diameter has to be less then $d = 20cm$ and at $f = 5MHz$ less then $d = 22cm$. The 5MHz plot shows two steps which are caused by the relay in parallel with the serial capacitor C_2 within the coupler. With the relay contact open, the efficiency is usually less than when closed. When the antenna diameter is between $9cm < d < 13cm$, the

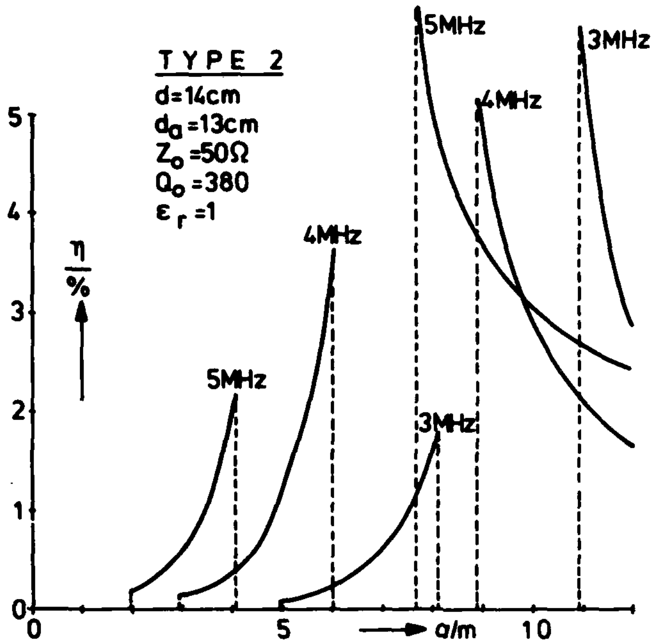


Fig. 43: Total efficiency of antenna type 2, versus length a of the antenna elements. Parameter: frequency f

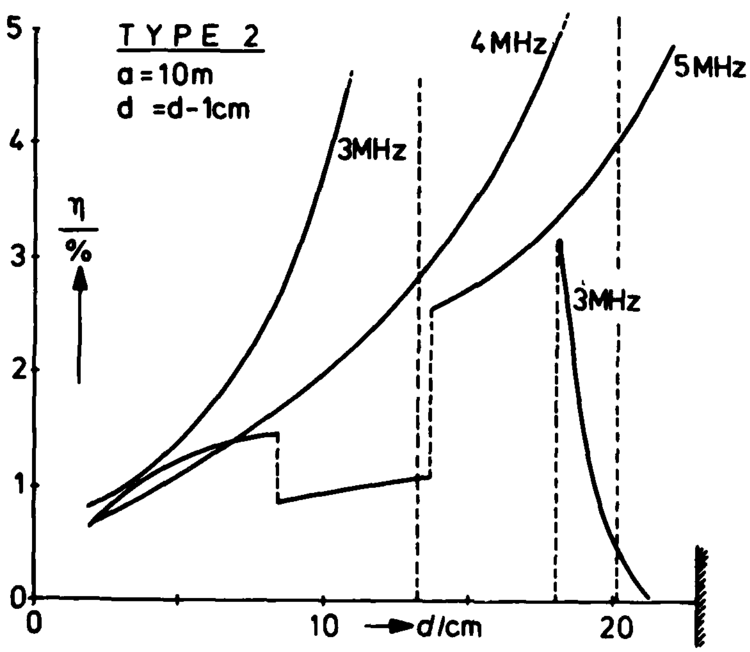


Fig. 44: Total efficiency of antenna type 2 versus diameter d of the antenna elements. Parameter: frequency f

antenna can only be tuned if C_2 is activated.

The characteristic impedance Z_o of the feeder line strongly influences the antenna efficiency as shown in Fig.45. At 3MHz the characteristic impedance has to be within 20 Ohms to 40 Ohms to get an efficiency betweeen 0.6% and 5.5% with the stated antenna dimensions. With any other characteristic impedance the antenna is not tunable. At 4MHz an efficiency of up to 9% can be achieved and the characteristic impedance has to be between 20 Ohms and 80 Ohms. At 5MHz the antenna is tunable for 20 Ohms$<Z_o<$120 Ohms. The maximum efficiency is 9.5%.

Also, for type 2 the optimum antenna has to be a compromise between high efficiency and tunability in the full frequency range.

An antenna with the dimensions:

$$a = 10m$$
$$d = 12.8cm$$
$$Z_o = 50\ Ohm$$
$$\varepsilon_r = 1.0$$

will give an excellent result of more than 11% total efficiency at f = 3MHz. It will be easily tunable at this frequency as shown in Fig.46. With an inductivity of L = 46µH the current in the coil will be very low and the Q-factor nearly as high as possible because all turns are used. Capacitor C_1 will be tuned to C_1 = 52pF and capacitor C_2 will be by-passed. The coil will convert 33% of the generator power to heat and only 0.16% of the variable capacitor. The losses of the coaxial feeder line are about 16% of the generator power and the contact losses and the losses of the antenna itself will be 40% of the available generator power. The losses of the antenna are uncharacteristicly high at this configuration. The reason is that the radiating gap between the antenna elements and the roof is very narrow. Therefore the radiation resistance is very low (10mΩ). The loss resistances are more than three times greater. On the other hand, this narrow gap causes a quite low imaginary part of the antenna impedance which is transformed by the coaxial line to a capacitive load of about 9pF, roughly equivalent to a 1m whip-antenna. but with significantly higher real part. As mentioned before, the total efficiency of this antenna is more than 11% at 3MHz, 14dB better than the existing antenna. For f = 4MHz and f = 5MHz the antenna efficiency is considerably smaller, as the current through the variable coil is higher and the number of turns used is very small. So the antenna will radiate about 2.5% of the available generator power at 4MHz and only 1% at 5MHz. The power lost in the variable coil will be 92% and 96%, respectively.

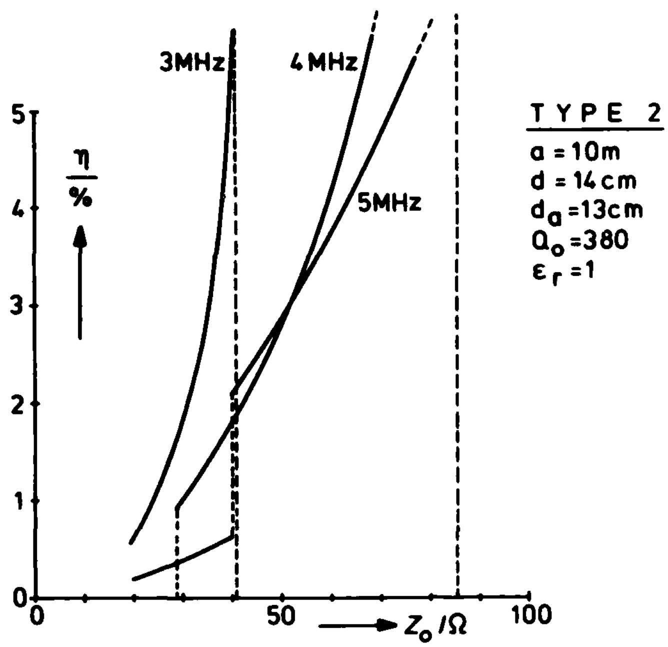


Fig. 45: Total efficiency of antenna type 2 versus characteristic impedance Z_o of the feeder line. Parameter: frequency f

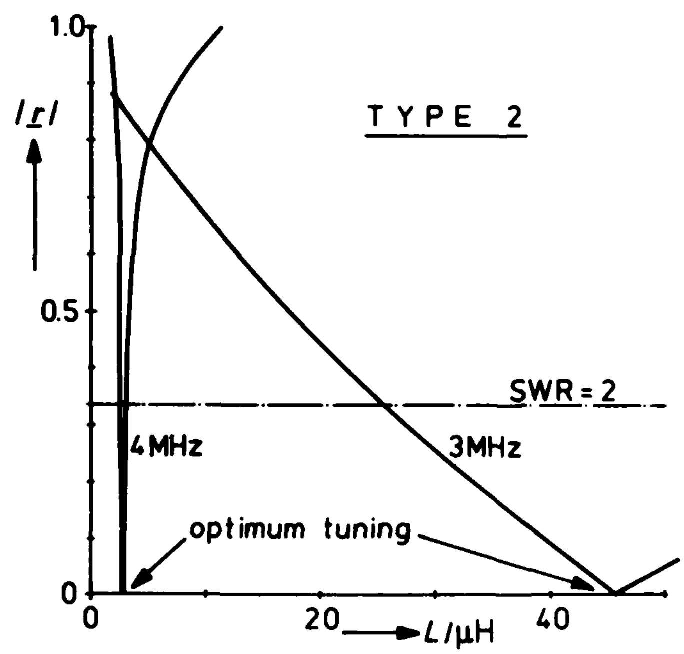


Fig. 46: Comparison of the dependence of the coupler input reflection coefficient r on the inductance L of the variable coil at 3 MHz and 4 MHz

Fig.47 shows the drawing of antenna type 2 with the proposed features.

7. Bandwidth of the tuned antenna

The bandwidth of electrical small antennas is usually small and the antenna's reactive power is much greater than the radiated power. For the antenna-coupler configuration, the bandwidth calculation needs more effort than for the antenna itself since the matching network is tunable and lossy.

The bandwidth has also been computed utilizing the program described before. When the coupler elements have been tuned for a fixed frequency, the input frequency is varied slightly, with fixed coupler elements. For each frequency the output power and the coupler input-reflection coefficient is calculated and plotted, as shown in Fig.48 for antenna type 1. The frequency variation is $\pm$10kHz and the centre frequencies are 3MHz, 4MHz and 5MHz.

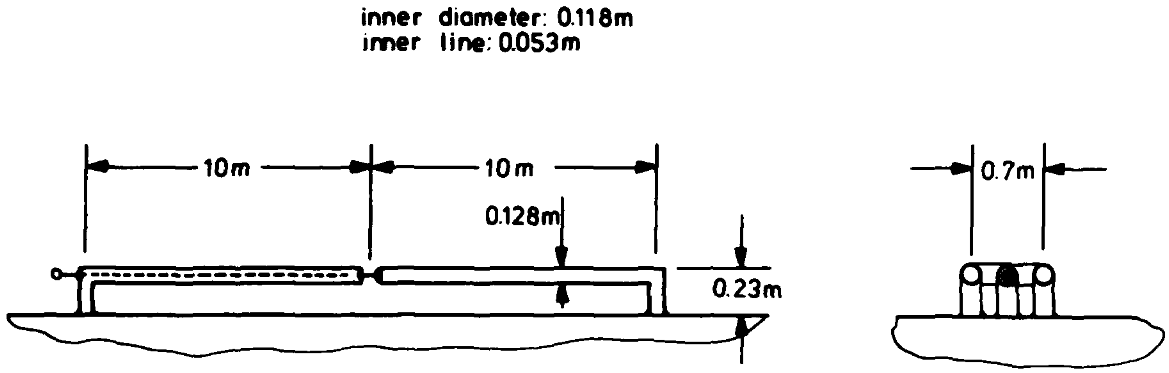
The output power is nearly independent of the frequency variation with this antenna type when the bandwidth necessary for single-sideband (SSB) communication (2.7kHz) is used. The output power changes less than 0.8% in this range and the reflection coefficient is less than 6%, corresponding to a standing-wave ratio of SWR $\leq$ 1.128. For telegraphy (CW) and for radio teletype (RTTY) communications the results will be even better because of the smaller bandwidth.

Fig.49 shows the corresponding plots for antenna type 2 with the optimized outlines (see chapt.6). The bandwidth achievable with this type of antenna is much smaller, but it will be sufficient even for SSB communication at the lowest and therefore most critical frequency of 3MHz ($B \approx$ 2.4kHz). The bandwidth tolerable from the maximum transmitter load standing-wave ratio of SWR $\leq$ 2 is distinctly lower ($\approx$ 1.36kHz at 3MHz). This bandwidth would be detracting for a long-time sinusoidal SSB modulation signal. As speech consists of a mixture of different frequencies at each time-interval, the maximum RMS standing-wave ratio will usually not be exeeded with a speech signal of 2.7kHz bandwidth.

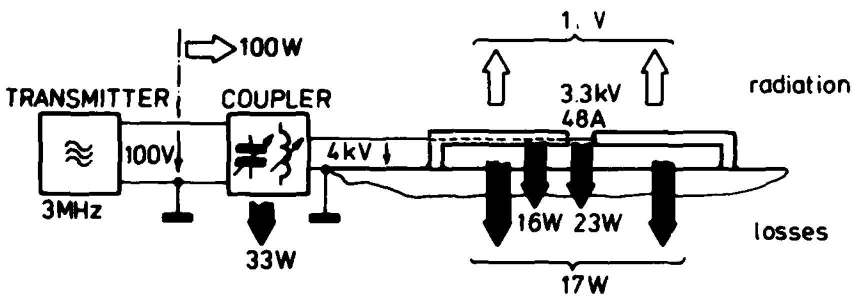
8. Measurement results

We investigated the currently existing antenna type 1 as well as a test antenna of type 2 with the non-optimum outlines:

$$a = 12\text{m}$$
$$d = 2\text{cm}$$



a) outline dimensions



b) flow graph for 100 W available transmitter power

Fig. 47: Antenna type 2, optimized for f = 3 MHz

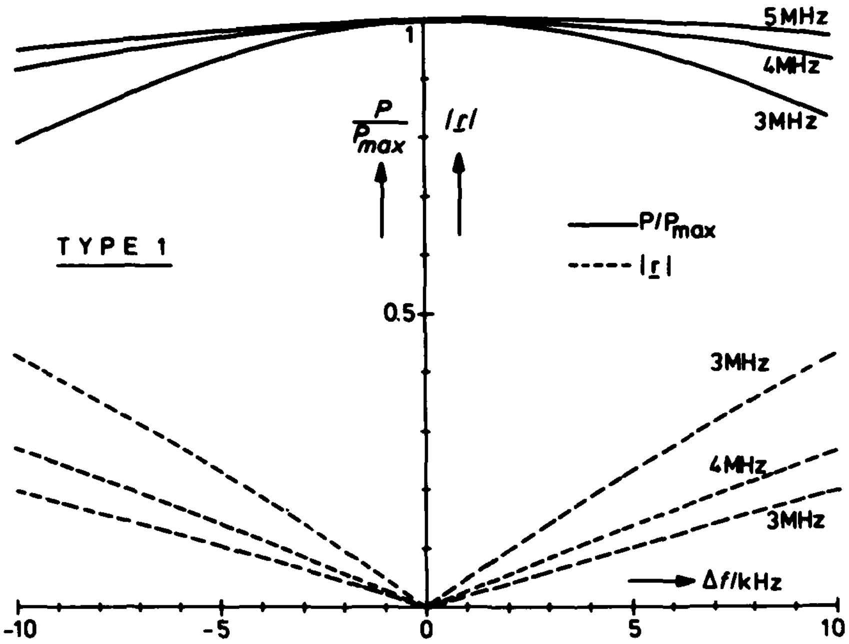


Fig. 48: Antenna type 1 - radiated power P and input reflection coefficient r versus frequency offset. Centre frequencies are 3 MHz, 4 MHz and 5 MHz

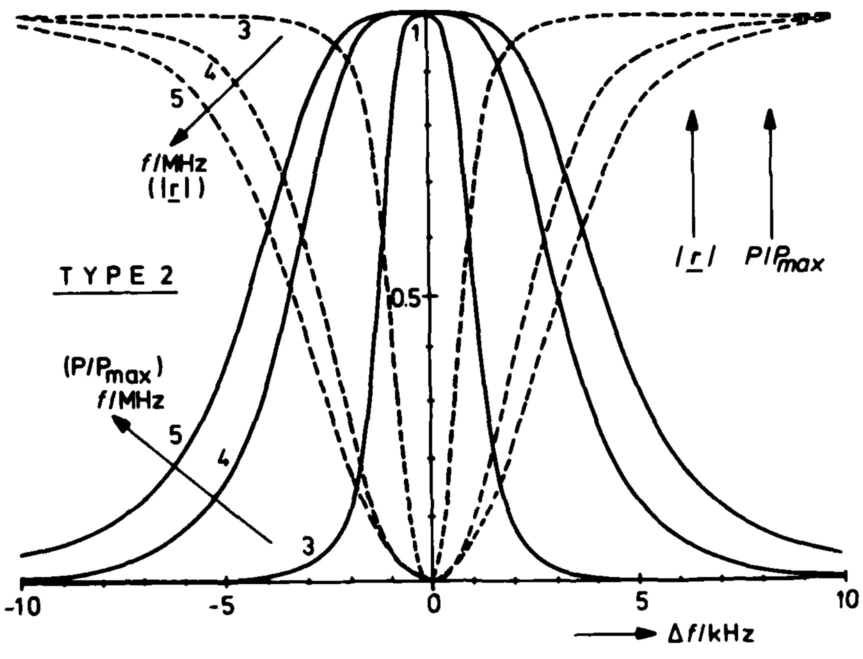


Fig. 49: Optimized antenna type 2 - radiated power and input reflection coefficient r versus frequency offset

$Z = 50$ Ohm
$\varepsilon_r = 1.2$
$D = 1.7$cm

The antennas were measured on top of the train.

Fig.50 shows the calculated and the measured imaginary part of the input impedance of antenna type 1, depending on the frequency. The real part of the impedance is much smaller than its imaginary part and cannot be measured directly with a vector impedance meter. At 3MHz the measured inductive impedance fits with the calculated value. At higher frequencies, the measured values are distinctly higher than the calculated values since the quarter-wavelength resonance appears at a lower frequency than expected from the antenna length (8.3MHz). Therefore, at the real antenna there has to be some additional capacitors. This result is quite reasonable since the capacitive insulator constructions that hold the antenna elements and the connection lines betweeen the elements have not been considered.

Fig.51 shows the imaginary part of the calculated and measured impedance of antenna type 2 at the coaxial feeder point. Again the real part is too small to be measured. The imaginary part is capacitive, since the inductive imaginary part of the antenna impedance at the symmetrical feeder point is transformed by the coaxial line, the length of which is close to the quarter wavelength.

The transformation is very sensitive to any change of the dimensions or the frequency since the antenna and line together act as a high-Q-resonance circuit close to the resonance. So, for example, the input reflection coefficient at the coaxial input shows a frequency dependence on the phase angle which is nearly four times greater than the frequency dependence on the phase angle of the antenna's reflection coefficient itself. Also, the small bandwith of antenna type 2 compared to type 1 is caused by the sensivity of this transformation.

The measured values of the coaxial input impedances imaginary part show an offset of about 25 Ohms compared to the calculated values. This effect is caused by the extremely long vertical short circuit rods (22cm) and by the actual shape of the roof, which for the calculations has been assumed to be flat. This assumption is not accurate if the distance between the fictitious line current and the roof becomes as high as with this test antenna.

Besides this offset, the frequency dependence of the measured and the calculated impedance correspond quite well.

Fig.52 shows the calculated and measured frequency dependence of

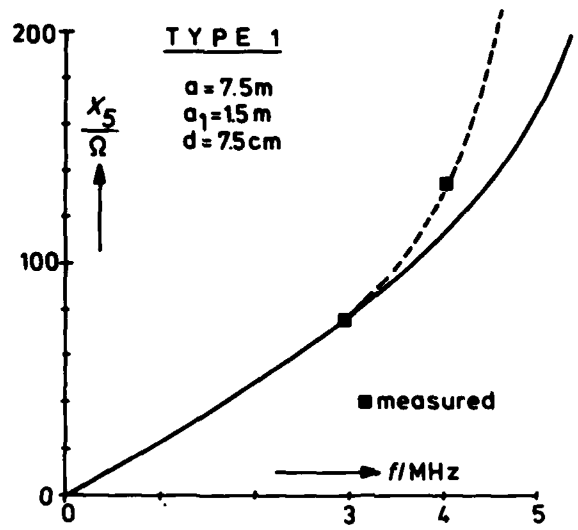


Fig. 50: Antenna type 1 - comparison of calculated and measured imaginary part X_5 of the antenna impedance $\underline{Z}_5$

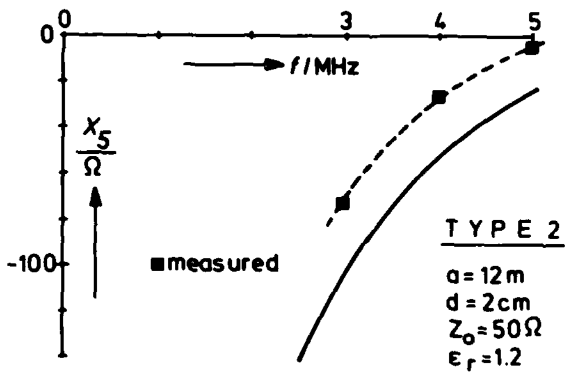


Fig. 51: Antenna type 2 - comparison of calculated and measured imaginary part X_5 of the antenna impedance $\underline{Z}_5$

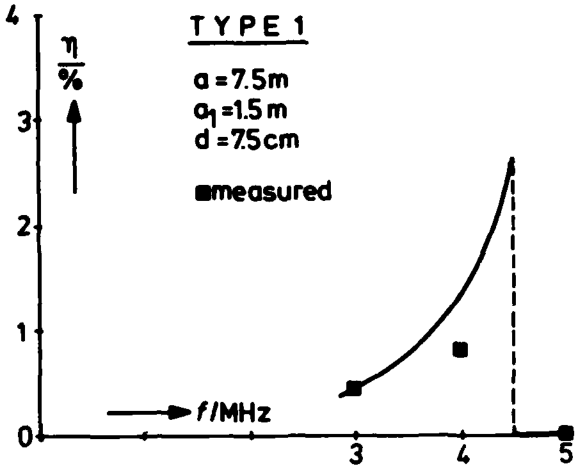


Fig. 52: Antenna type 1 - calculated and measured total efficiency η versus frequency

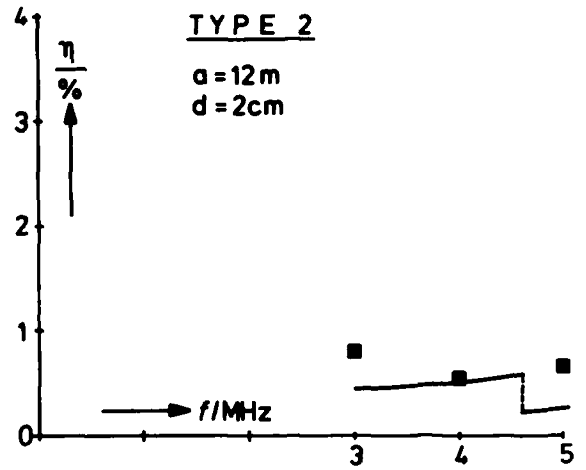


Fig. 53: Antenna type 2 - calculated and measured total efficiency versus frequency

the total efficiency of antenna type 1 and coupler. At 3MHz the measured and the calculated results are identical, like the measured and calculated results of the antenna impedance at this frequency. At 4MHz the measured result is about 3dB worse than the calculated value. At frequencies greater than about 4.5MHz the antenna is theoretically not tunable. This is confirmed by the measured results.

The correspondence of calculated and measured data is quite good, although several estimations and some approximations have been used. Besides this, the train could not be measured standing at a free area. Furthermore, the uncertainity of the field strength measurement setup is not better than about 2dB.

The correspondence of measured and calculated data of the total efficiency of antenna type 2, shown in Fig.53, is also quite good. The difference at 3MHz is about 3dB, at 4MHz less than 0.3dB and at 5MHz about 4dB.

Fig. 55 shows the comparison of the measured and the calculated radiation patterns in the plane $\theta = 90^{\circ}$. The investigated antenna is antenna type 2 at a frequency of f = 3MHz. The receiving antenna used vertical polarization. The distance between train and receiver was about 1km. The test car was equipped with an active loop antenna (R&S HFH2-Z2) and a receiver (R&S ESH2) for the field strength measurements (see Fig.59). *The test points* (C1 - C4) which were chosen are shown at the map of Fig.54. Within 2dB the calculated curve fits with the measurement results.

To document how the electrical field strength depends on the distance, compared to the CCIR-Recommendation 368-4, Vol. V, Geneve 1982 the field strength was measured at three more testpoints (MP2, MP3 and MP4) at distances of 2km, 3.3km and 4.6km. The azimuth angles were between 76° and 83°. Referred to an effective radiated power of 1kW, the measurement results for 3MHz (Fig.56) and 5MHz (Fig.57) are within the CCIR-curves for ground-conductivities of $\sigma = 3 \cdot 10^{-2}$S/m and $\sigma = 10^{-5}$S/m and very close to the documented $\sigma = 10^{-3}$S/m-curve (Europe near the alpes: $\sigma = 3 \cdot 10^{-3}$S/m).

With the results of the measurements and the calculations and with the help of the CCIR-Recomendations and Reports it is therefore possible to estimate the communication distance for ground-wave propagation, achievable with the system.

Fig. 58 shows the flow graph from the transmitter T to the receiver R. The transmitter delivers its available power P_a to the matching network NW_T with the efficiency η_T . The network output power $\eta_T \cdot P_a$ is radiated from the antenna to free space with the partial vertical directivity d_T in the direction of the

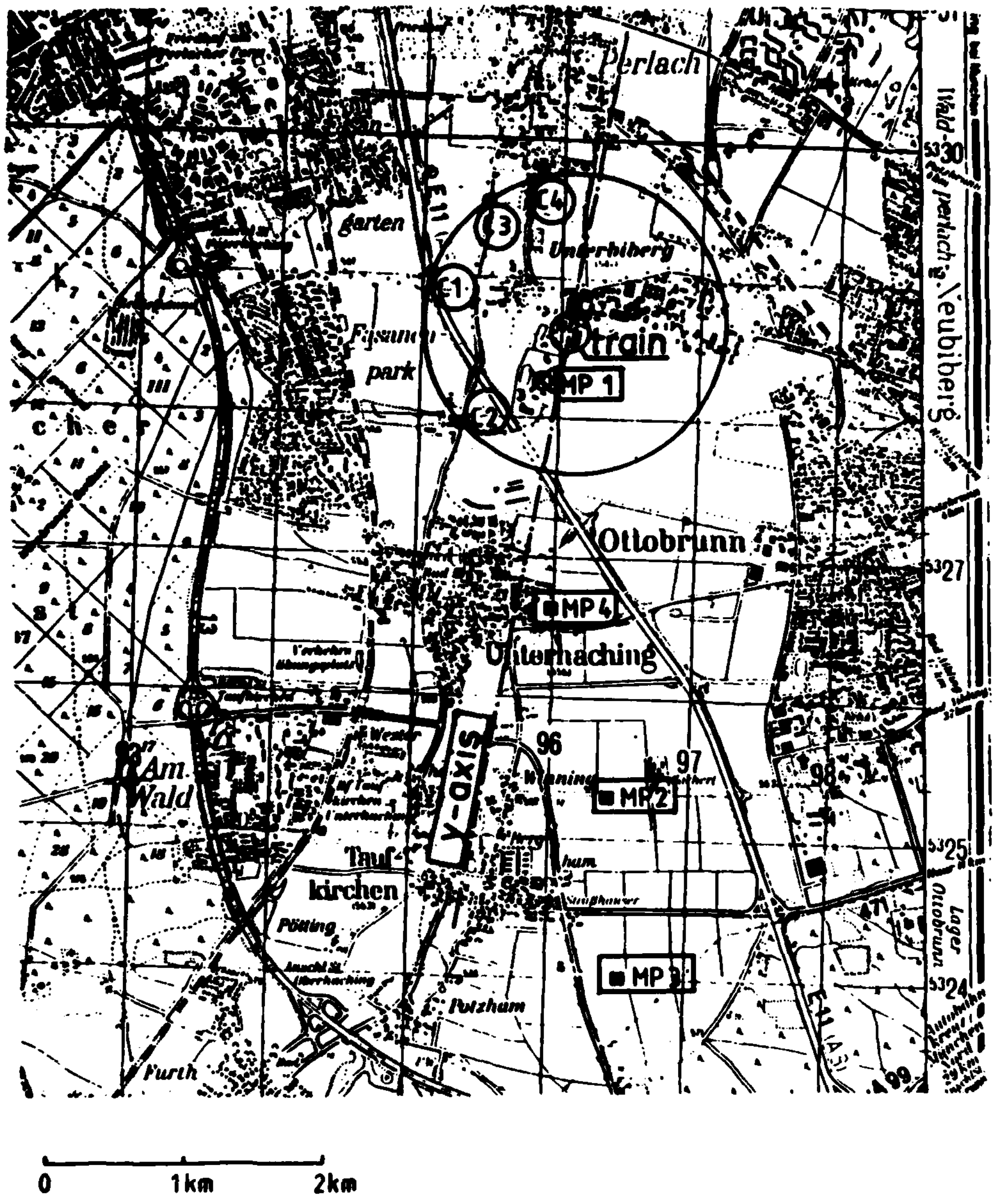


Fig. 54: Location of test points for the measurement of

- total efficiency η (MP1)
- propagation (MP2-MP4)
- radiation pattern (C1 - C4)

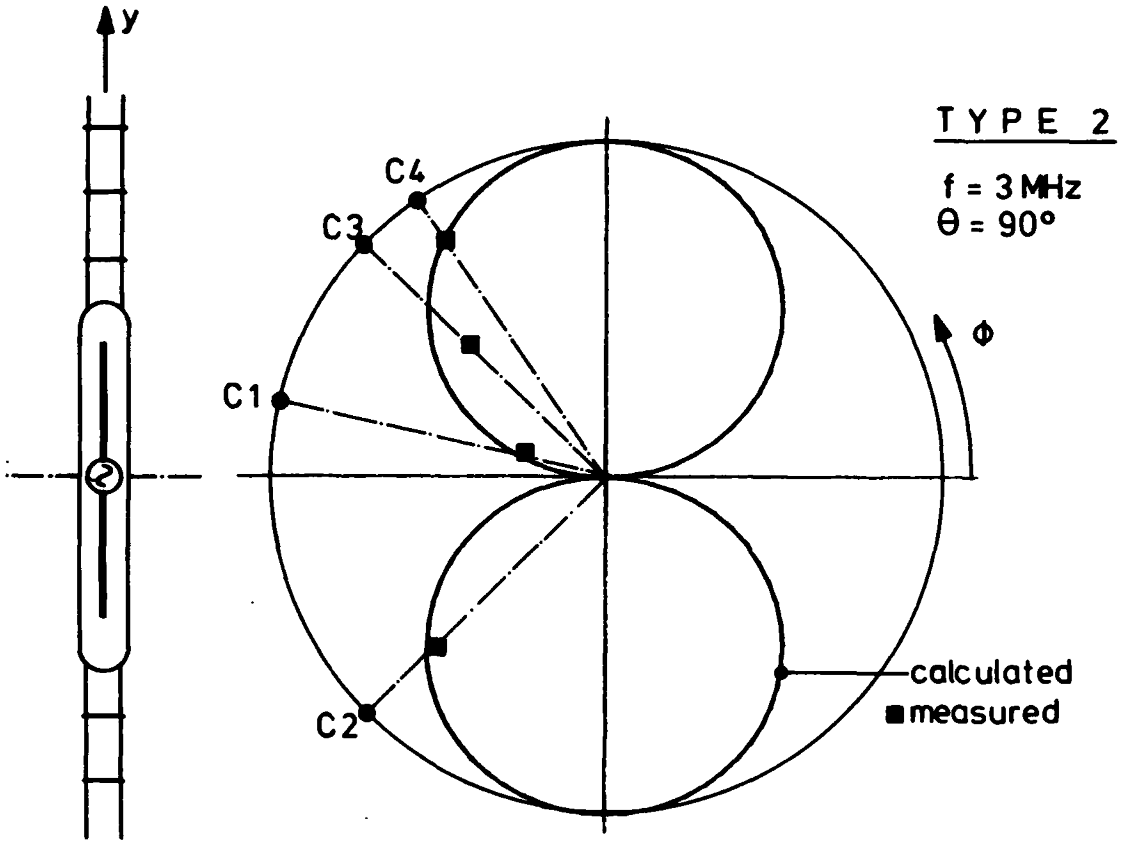


Fig. 55: Antenna type 2 - comparison of calculated and measured radiation pattern in the $\Theta = 90^{\circ}$-plane

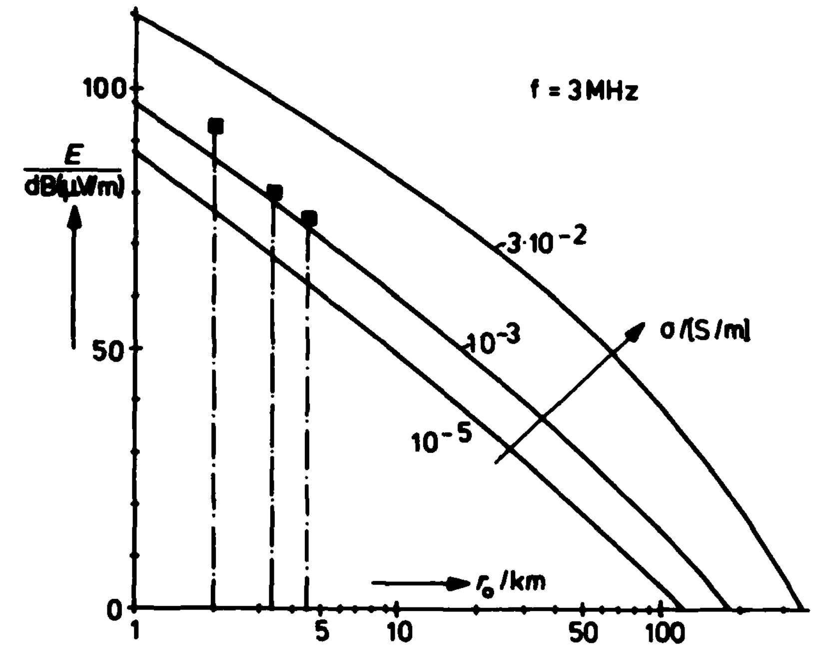


Fig. 56: Antenna type 1 - comparison of measured fieldstrength and ground-wave propagation curves with different ground conductivity σ (P_r = 1kW)

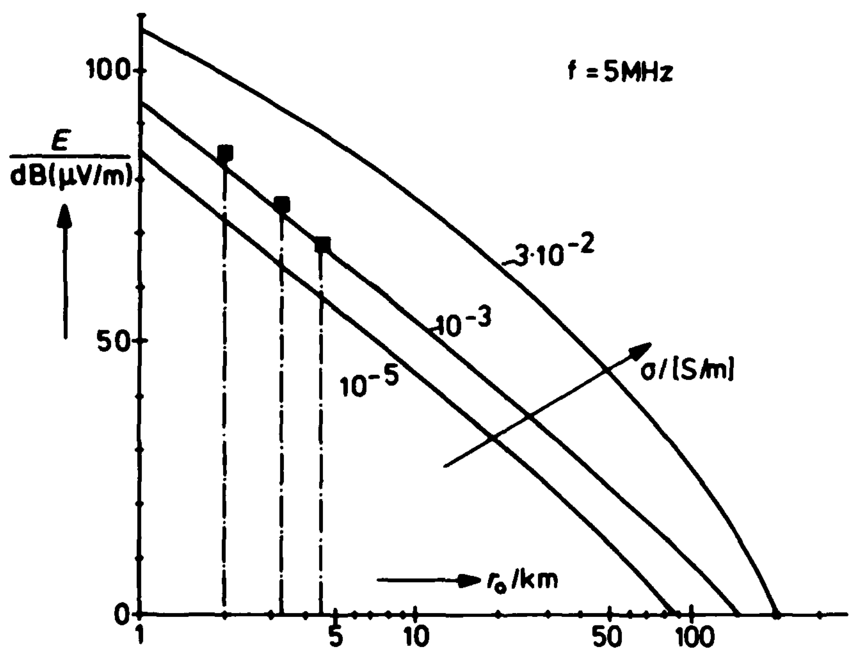


Fig. 57: Antenna type 1 - comparison of measured fieldstrength and ground-wave propagation curves with different ground conductivity σ (P_r = 1kW)

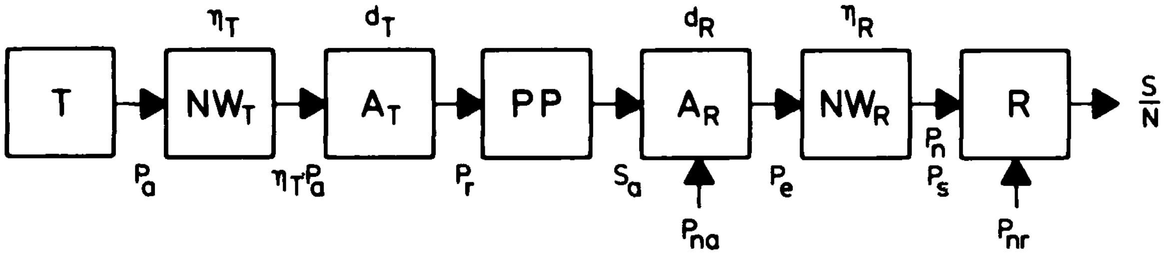


Fig. 58: Flow chart for signal and noise in a radio telecommunication link

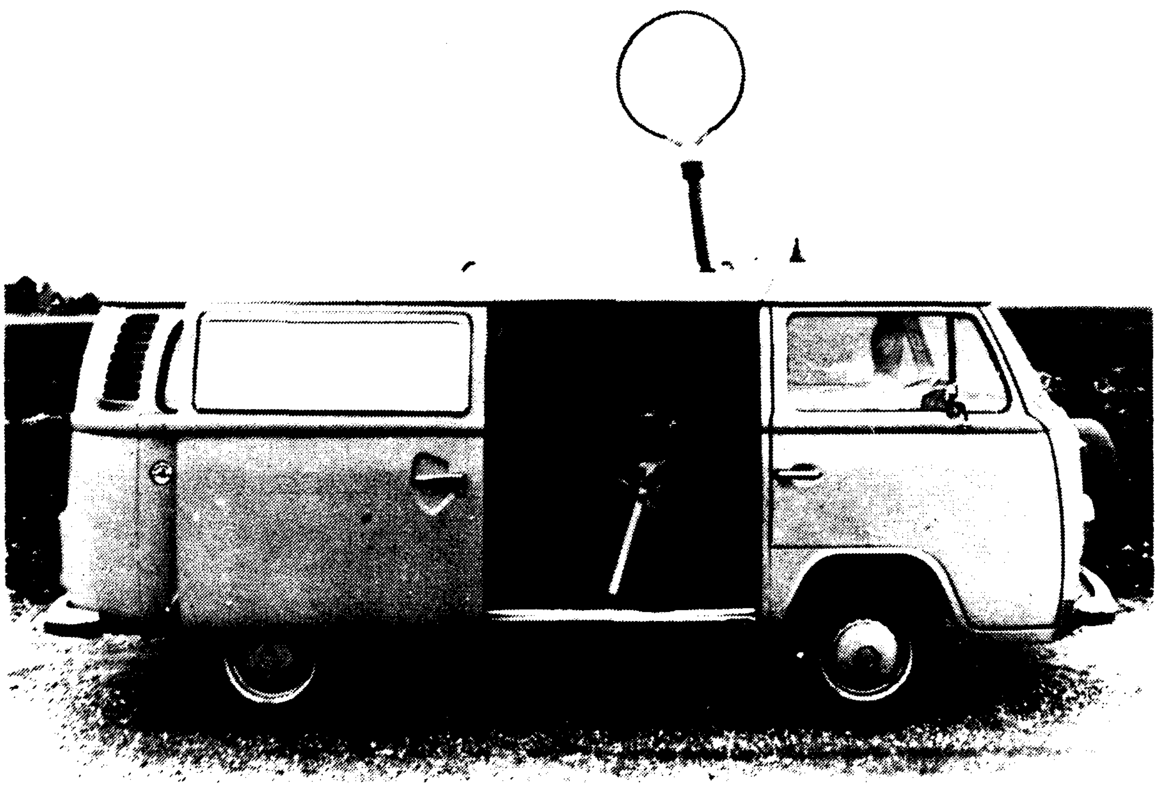


Fig. 59: Test car, equipped with an active loop-antenna (HFH2-Z2, R&S) and a selective level meter (ESH2, R&S)

receiver. The propagation path PP is described by the CCIR-recommendation 368-4. The electrical field strength E at the receiving antenna can be found by help of this documentation depending on the frequency, the radiated power, the distance, and the ground conductivity.

The receiving antenna A delivers the power

$$P_e = \frac{\lambda_o^2}{16 \cdot \pi} \cdot d_R \cdot \frac{E^2}{Z_i} \tag{60}$$

to the matching network. The partial directivity of the receiving antenna is d_R. The matching network has the efficiency η_R.

In additional to the signal field strength E there will be a noise field strength e_n at the receiving antenna which depends not only on the frequency, but also on the surroundings ("man-made" noise sources) and the time of day. The noise power P_{na} received by the antenna inside the receiver bandwidth B can be added to the available thermal noise power P_{no} of the antenna source resistance. Because of this, the noise figure F_a of the antenna will be:

$$F_a = \frac{P_{na}}{P_{no}} \tag{61}$$

$$P_{no} = k \cdot T_o \cdot B \tag{62}$$

The typical noise figure of antennas depending on the frequency and the surrounding is documented in Report 670 of the CCIR-Recommendations and Reports, Vol.I. In the frequency range from 3MHz to 5MHz the noise figure is stated to be between F_a= 40dB ("silent surrounding") and F_a= 70dB.

Signal and noise are attenuated by the lossy antenna matching network NW_R of the receiving system. The signal P_S, delivered to the receiver is:

$$P_s = \eta_R \cdot P_e \tag{63}$$

The noise power P_n, delivered to the receiver is:

$$P_n = (P_{na} + P_{no}) \cdot \eta_R \tag{64}$$

As the receiver itself produces noise, the signal-to-noise ratio is decreased further:

$$\frac{S}{N} = \frac{P_s}{P_n + P_{nr}}$$

The additional receiver noise P_{nr} is usually described by the receiver noise figure F_r:

$$F_r = \frac{P_{nr}}{P_{nO}} \tag{65}$$

where P_{nO} is the noise power delivered by the source resistance at the temperature T_O (normally: T_O = 288K). The receiver noise figure in the frequency range from 3MHz to 5MHZ is typically about $F_{r,dB} \approx 8dB$. As the antenna is matched to the receiver input impedance, there is no difference between available and delivered power.

The signal-to-noise ratio at the receiver output port thus becomes:

$$\frac{S}{N} = \frac{\eta_R \cdot \lambda_o^2 \cdot d_R \cdot E^2}{16 \cdot \pi \cdot Z_i \cdot k \cdot T_O \cdot B \cdot [\eta_R(F_a+1) + F_r]} \tag{67}$$

For a certain desired minimum signal-to-noise ratio $(S/N)_{min}$ the electrical field strength E at the receiving antenna must not be lower than:

$$\frac{E}{[dB(\mu V/m)]} = 20 \cdot \log \frac{f}{[MHz]} + (\frac{S}{N})_{min} - 64 +$$

$$+ 10 \cdot \log \frac{B}{[kHz]} + 10 \cdot \log(1 + F_a + \frac{F_r}{\eta_R}) \tag{68}$$

By help of the CCIR-field strength curves the maximum usable distance can be estimated, using this equation.

A typical example is:

- SSB-communication (B = 2.7kHz)
- good readability ($(S/N)_{min}$ = 20dB)
- antenna noise figure F_a = 40dB
- receiver noise figure F_r = 8dB
- available transmitter power P_a = 300W
- medium dry ground ($\sigma = 10^{-3}$S/m)

If transmitter and receiver use the railroad-train antenna type 1 the maximum usable distance for ground-wave communication will be about 30km at 3MHz.

If a better antenna, like a 15ft-whip antenna or the optimized antenna type 2 is used for transmitting only, the maximum distance will increase to 70km at 3MHz.

Even with a poor total antenna efficiency, the antenna noise is much stronger than the receiver noise and the influence of different antenna types on the receiver sensitivity is extremly low (less than 1dB). Therefore the maximum usable distance can

only be increased by increasing the radiated signal power or by decreasing the signal bandwidth or the minimum signal-to-noise ratio.

If, for example, a minimum S/N-ratio of $(S/N)_{min}$ = 10dB will be accepted and RTTY with a frequency shift of 700Hz is used, the maximum usable distance at 3MHz will be 100km with antenna type 1 and close to 200km with the optimized antenna type 2.

The good conductivity of the rails and the influence of the overhead-contact lines will increase the maximum distance further.

Using the sky-wave for communication will change the attenuation of the propagation path remarkably. In this case the maximum usable distance depends additionally on the time of day, the season, and the sun-spot activities. The maximum distance can be much greater than that achievable with the ground-wave. The probability that ionospheric propagation can be used for communication can be taken from actual prediction charts.

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A P P E N D I X   B

Computer Programs

```
10    ! ****************************************************************
20    ! *                                                              *
30    ! *       CALCULATION OF THE RADIATION PATTERN (TYPE-1)          *
40    ! *                                                              *
50    ! *           (written for Hewlett Packard 9845B)                *
60    ! *                                                              *
70    ! ****************************************************************
80    !
90    !
100   !
110   OPTION BASE 1
120   A=7.5                         !element length [m]
130   H=.23                         !actual height of line current [m]
140   F=3E6                         !frequency [Hz]
150   C=3E8                         !light velocity [m/s]
160   DEG
170   La=C/F                        !lambda [m]
180   B=360/La                      !beta [degree/m]
190   K=20                          !element intersection
200   I0=1                          !feeder current [A]
210   Im=I0/COS(B*A)                !maximum current
220   Hm=H/La*Im                    !maximum magnetic fieldstrength
230   DIM Hf(3,19,19)
240   PLOTTER IS "GRAPHICS"
250   GCLEAR
260   GRAPHICS
270   SCALE 0,19,0,Hm
280   M=0
290   LORG 5
300   FOR Phi=0 TO 90 STEP 5        !azimuth steps
310   GCLEAR
320   AXES 1,Hm/10
330   MOVE 10,Hm*.8
340   LABEL "Phi=";Phi;"Grad"
350   M=M+1
360   N=0
370   FOR Th=0 TO 90 STEP 5         !elevation steps
380   N=N+1
390   H1=0
400   H2=0
410   H3=0
420   H4=0
430   Dl=A/K
440   FOR I=0 TO K                  !sum of complex magnetic field
450   Y=I*Dl                        !strengths for each phi and theta
```

```
460   K1=COS(B*(A-Y))
470   D=Y*SIN(Th)*SIN(Phi)
480   D1=D+H*COS(Th)
490   D2=D-H*COS(Th)
500   H1=H1+K1*COS(B*D1)
510   H2=H2+K1*SIN(B*D1)
520   H3=H3-K1*COS(B*D2)
530   H4=H4-K1*SIN(B*D2)
540   NEXT I
550   Hn=D1/2/La*Im
560   Hf(1,N,M)=D1/2/La*Im*ABS(COS(Phi))*SQR((H1+H3)^2+(H2+H4)^2)
570   Hvr=H/La*Im*SIN(Th)*COS(B*A*SIN(Th)*SIN(Phi))
580   Hvi=H/La*Im*SIN(Th)*SIN(B*A*SIN(Th)*SIN(Phi))
590   Hvr=Hvr-H/La*SIN(Th)
600   Hhr=D1/2/La*Im*COS(Th)*SIN(Phi)*(H1+H3)
610   Hhi=D1/2/La*Im*COS(Th)*SIN(Phi)*(H2+H4)
620   Hf(2,N,M)=SQR((Hvr+Hhr)^2+(Hvi+Hhi)^2)
630   Hf(3,N,M)=SQR(Hf(1,N,M)^2+Hf(2,N,M)^2)
640   MOVE N,Hf(1,N,M)
650   LABEL "*"
660   MOVE N,Hf(2,N,M)
670   LABEL "+"
680   NEXT Th
690   NEXT Phi
700   !
710   ! ********************************************************
720   !
730   CREATE "AntDa",60            !store data on tape
740   ASSIGN #1 TO "AntDa"
750   MAT PRINT #1;Hf
760   ASSIGN * TO #1
770   END
```

```
10    ! ************************************************************
20    ! *                                                          *
30    ! *     CALCULATION OF THE RADIATION PATTERN (TYPE-2)        *
40    ! *                                                          *
50    ! ************************************************************
60    !
70    !
80    !
90    OPTION BASE 1
100   A=12                          !element length [m]
110   H=.23                         !actual height of line current [m]
120   F=3E6                         !frequency [Hz]
130   C=3E8                         !light velocity [m/s]
140   DEG
150   La=C/F                        !lambda [m]
160   B=360/La                      !beta [degree/m]
170   K=20                          !element intersection
180   I0=1                          !feeder current [A]
190   Im=I0/COS(B*A)                !maximum current
200   Hm=2*H/La*Im
210   DIM Hf(3,19,19)
220   PLOTTER IS "GRAPHICS"
230   GCLEAR
240   GRAPHICS
250   SCALE 0,19,0,Hm
260   M=0
270   LORG 5
280   ! -------------------------------------------------------------
290    FOR Phi=0 TO 90 STEP 5       !azimuth steps
300    GCLEAR
310    AXES 1,Hm/10
320    MOVE 10,Hm*.8
330    LABEL "Phi=";Phi;"Grad"
340    M=M+1
350    N=0
360    ! ------------------------------------------------------------
370     FOR Th=0 TO 90 STEP 5       !elevation steps
380     N=N+1
390     Hl=0
400     Dl=A/K*2
410    ! ------------------------------------------------------------
420      FOR I=0 TO K               !vector sum of magnetic
430      Y=I*Dl-A                   !fieldstrengths
440      Kl=COS(B*(A-ABS(Y)))
450      D=Y*SIN(Th)*SIN(Phi)+H*COS(Th)
460      Hl=Hl+Kl*SIN(B*D)
470      NEXT I
480    ! ------------------------------------------------------------
```

```
490      Hf(1,N,M)=Dl/La*Im*ABS(COS(Phi)*Hl)
500      Hf(2,N,M)=Dl/La*Im*ABS(SIN(Phi)*COS(Th)*Hl)+
         H/La*Im*2*ABS(SIN(Th)*SIN(B* A*SIN(Th)*SIN(Phi)))
510      Hf(3,N,M)=SQR(Hf(1,N,M)^2+Hf(2,N,M)^2)
520      MOVE N,Hf(1,N,M)
530      LABEL "*"
540      MOVE N,Hf(2,N,M)
550      LABEL "+"
560      NEXT Th
570     ! ------------------------------------------------------------------
580     NEXT Phi
590    ! --------------------------------------------------------------------
600    CREATE "AntDn",60                          !store data on tape
610    ASSIGN #1 TO "AntDn"
620    MAT PRINT #1;Hf
630    ASSIGN * TO #1
640    END
```

```
10    ! ***************************************************
20    ! *                                                 *
30    ! *     CALCULATION OF THE RADIATION RESISTANCE     *
40    ! *                                                 *
50    ! *       (Needs data from radiation pattern)       *
60    ! *                                                 *
70    ! ***************************************************
80    !
90    !
100   OPTION BASE 1
110   DIM H(3,19,19)
120   ASSIGN #1 TO "AntDa"
130   MAT READ #1;H               !reading data from tape
140   ASSIGN * TO #1
150   RAD
160   D=5*PI/180
170   Z=120*PI                    !intrinsic impedance
180   Rs=0
190   !
200   ! *******  numeric integration of the power density *****
210   ! -------------------------------------------------------
220    FOR M=1 TO 19
230    R=0
240   ! -------------------------------------------------------
250     FOR N=1 TO 19
260     R=R+H(3,N,M)^2*SIN((N-1)*D)*D
270     NEXT N
280   ! -------------------------------------------------------
290    Rs=Rs+R*D
300    NEXT M
310   ! -------------------------------------------------------
320   Rs=Rs*4*Z
330   PRINTER IS 0
340   PRINT "Radiation resistance Rs =";Rs;" Ohm"
350   END
```

```
10    ! ***********************************************************************
20    ! *                                                                     *
30    ! *              CALCULATION OF THE TOTAL ANTENNA EFFICIENCY            *
40    ! *                                                                     *
50    ! *                          (Antenna type 2)                           *
60    ! *                                                                     *
70    ! *                                                                     *
80    ! ***********************************************************************
90    !
100   !
110   !
120   !
130   PRINTER IS 16
140   FIXED 2
150   PRINT PAGE
160   OPTION BASE 1
170   DIM L(101),R(101),C(101)
180   Clmin=1.2E-11     !  Couplercapacity Cl minimum/[F]
190   Clmax=5.00E-10    !  Couplercapacity Cl maximum/[F]
200   C22=2.00E-10      !  Couplercapacity C2 /[F]
210   Lmin=1E-6         !  Couplerinductivity minimum/[H]
220   Lmax=5.0E-5       !  Couplerinductivity maximum/[H]
230   Scu=5.8E7         !  Conductivity of Cu/[S/m]
240   Sal=3.0E7         !  Conductivity of Al/[S/m]
250   Hmax=.23          !  Maximum antenna height/[m]
260   A=12              !  Element length/[m]
270   D=2.0E-2          !  Antenna diameter/[m]
280   C=3E8
290   MO=4*PI*1E-7      !  Permittivity
300   Er=1.2            !  Epsilon(rel.) of coaxial feeder line
310   Z1=50             !  characteristic impedance of feeder line/[Ohms]
320   Da=1.7E-2         !  Diameter of outer conductor of the  feeder line/[m]
330   Di=Da*EXP(-Z1*SQR(Er)/60)
340   Lt=A              !  Length of coaxial line/[m]
350   Rq=50             !  Source resistance/[Ohm]
360   P0=100            !  Available generator power/[W]
370   !
380   ! ************************************ contact resistances*****************
390   !
400   Rc1=2.0E-2
410   Rc2=2.0E-2
420   Rc5=2.0E-2
430   Rc6=2.0E-2
440   !
450   ! ************************************** Transformer T **********************
460   !
470   Rsoll=Rq/4
480   RAD
490   !
500   ! *************************************** Q-factor of the variable coil *****
*
510   !
```

```
520   Q1=380
530   INPUT "frequency/[MHz]: ",F
540   IF (F<6) AND (F>2) THEN 590
550   DISP "frequency not correct!"
560   BEEP
570   WAIT 2000
580   GOTO 530
590   F=F*1E6
600   Om=2*PI*F
610   Q0=Q1*SQR(F/5E6)                       ! actual Q-factor
620   B0=-1/Om/Lmax
630   H=SQR(Hmax*(Hmax-D))                   ! height of line current
640   H1=Hmax-D/2                            ! centre of antenna element
650   La=C/F                                 ! wavelength
660   Be=2*PI/La                             ! phase velocity
670   P=Be*A
680   W=2*P
690   Rv=A/D/3*SQR(F*M0/PI/Sal)*H1/H*(W+SIN(W))/W/COS(P)^2    ! loss resistance
700   Xa=60*LOG(2*(H1+H)/D)*TAN(P)           ! imaginary part of antenna impedance
710   Z0=120*PI
720   Rs=1.3*16*PI/3*Z0*(H/La*TAN(P))^2      ! approximated radiation resistance
730   Bee=Be/SQR(Er)                         ! phase velocity inside feeder line
740   L0=ATN(Xa/Z1)/Bee                      ! fictitious length
750   !
760   ! *********************************** losses of feeder line *************
770   !
780   Ro=SQR(F*M0/Scu/PI)*(1/Di+1/Da)
790   Rvc=Ro/4/Bee/COS(Bee*L0)^2*(2*Bee*Lt-SIN(2*Bee*L0)+SIN(2*Bee*(L0+Lt)))
800   !
810   ! *********************************** antenna impedance including losses **
820   !
830   R6=Rvc+Rs+Rv+Rc5
840   X6=Xa
850   ! ************************************************************************
860   CALL Div(1,0,R6,X6,G6,B6)              ! antenna admittance
870   T=TAN(Bee*Lt)
880   ! *********************************** transformation of feeder line *******
890   CALL Mul(0,T,R6/Z1,X6/Z1,Nr,Ni)
900   CALL Div(R6,X6+Z1*T,1+Nr,Ni,R5,X5)
910   ! ************************************************************************
920   R5=R5+Rc5                              ! contact resistance coupler - feeder
930   CALL Div(1,0,R5,X5,G5,B5)
940   PRINTER IS 0
950   PRINT "------------------------------------------------------------------
-----------"
960   PRINT "  INPUT DATA:     f = ";F*1E-6;" MHz    a = ";A;" m       d = ";D*100;
" cm"
970   PRINT "                  max. Q-factor: Q0 = ";Q0;"  when: L = Lmax"
980   PRINT "------------------------------------------------------------------
-----------"
990   PRINT "height of line current:                h = ";H*100;" cm"
1000  PRINT "radiation resistance:                  Rs = ";INT(1000*Rs+.5);" mOhm
```

```
"
1010  PRINT "loss resistance of antenna:          Rv = ";INT(1000*Rv+.5);" mOhm
"
1020  PRINT "loss resistance of feeder line:      Rl = ";INT(1000*Rvc+.5);" mOh
m"
1030  PRINT "imaginary part of antenna impedance: Xa = ";INT(10*Xa+.5)/10;" Ohm
"
1040  PRINT "coaxial imput impedance:             R5 = ";INT(1000*R5+.5)/1000;"
 Ohm"
1050  PRINT "                                     X5 = ";INT(1000*X5+.5)/1000;"
 Ohm"
1060  PRINT "contact resistances (Cl,C2,ant.-coax,coax-coupler: ";Rcl*1000;" |mO
hm]"
1070  PRINT "-----------------------------------------------------------------------
-----------"
1080  PRINT LIN(3)
1090  Durchl=0
1100  !
1110  ! ************************** relay contact par. C2 open or closed ********
1120  !
1130     FOR S2=1 TO 2              ! open: S2=2,  closed: S2=1
1140     ON S2 GOTO 1150,1170
1150     X45=0
1160     GOTO 1180
1170     X45=-1/Om/C22
1180     R4=R5+Rc2                  ! contact resistance at C2 or relay
1190     X4=X5+X45
1200     CALL Div(1,0,R4,X4,G4,B4)  ! admittance
1210     Z=0
1220     !
1230     ! ************************ rough tuning of inductance L ***************
1240     !
1250        FOR I=0 TO 100
1260        L=Lmin+I*(Lmax-Lmin)/100
1270        Bl=-1/Om/L
1280        Gl=-Bl/Q0*SQR(Bl/B0)     ! loss conductance of inductance L
1290        CALL Refl(G4,B4,Gl,Bl,Rb,Cl,Clmin,Clmax,Rcl,Rsoll,Om)
1300        IF Rb>.995 THEN 1350
1310        Z=Z+1
1320        R(Z)=Rb
1330        L(Z)=L
1340        C(Z)=Cl
1350        NEXT I
1360      ! *******************************************************************
1370      !
1380     ! ****************************** supervision of tuning ****************
*
1390     !
1400     PLOTTER IS "GRAPHICS"
1410     GCLEAR
1420     GRAPHICS
1430     SCALE -5,59,-.095,1.05
```

```
1440     STANDARD
1450     CSIZE 2.8
1460     AXES 10,.1
1470     LORG 5
1480  ! ------------------------------------------------------------------------
-
1490        FOR I=10 TO 50 STEP 10
1500        MOVE I,-.05
1510        LABEL I
1520        NEXT I
1530  ! ------------------------------------------------------------------------
1540        FOR I=1 TO 10
1550        MOVE -2.5,I/10
1560        LABEL I/10
1570        NEXT I
1580  ! ------------------------------------------------------------------------
1590     MOVE 3,.75
1600     CSIZE 3.3
1610     LABEL ":r:"
1620     MOVE 3,.7
1630     LABEL "^"
1640     MOVE 3,.68
1650     LABEL ":"
1660     MOVE 3,.66
1670     LABEL ":"
1680     MOVE 45,.04
1690     LABEL "--> L/uH"
1700     CSIZE 2.8
1710     MOVE 45,.36
1720     LABEL "**SWR=2**"
1730  ! ------------------------------------------------------------------------
1740        FOR I=1 TO Z
1750        MOVE L(I)*1E6,R(I)
1760        LABEL "+"
1770        NEXT I
1780  ! ------------------------------------------------------------------------
1790     LINE TYPE 4
1800     MOVE 1,1
1810     DRAW 50,1
1820     MOVE 50,0
1830     DRAW 50,1
1840     MOVE 1,0
1850     DRAW 1,1
1860     LINE TYPE 6
1870     MOVE 1,1/3
1880     DRAW 50,1/3
1890     LINE TYPE 1
1900     MOVE 40,1.025
1910     ON S2 GOTO 1920,1940
1920     LABEL "C2: short-circuit"
1930     GOTO 1950
1940     LABEL "C2 = 200 pF"
```

```
1950    Minr=1
1960    IF Z=0 THEN 3430
1970  !
1980  ! **************************************** fine tuning of inductance L *****
1990  !
2000       FOR I=1 TO Z                         ! searching minimum SWR
2010       IF R(I)>Minr THEN 2050
2020       Minr=R(I)
2030       Lminr=L(I)
2040       Minc=C(I)
2050       NEXT I
2060  ! ---------------------------------------------------------------------------
2070    Step=0
2080    IF Minr<.001 THEN 2350                  ! tuning successfull?
2090    Step=Step+1
2100    IF Step>20 THEN 3430
2110    IF (Step=20) AND (Minr<=1/3) THEN 2350   ! tuning acceptable?
2120    K1=1E-6
2130    U1=Lminr-K1/Step
2140    O1=Lminr+K1/Step
2150    IF U1>Lmin THEN GOTO 2170
2160    U1=Lmin                                 ! lower limitation
2170    IF O1<Lmax THEN GOTO 2190
2180    O1=Lmax                                 ! upper limitation
2190    St=2*K1/Step
2200  ! ---------------------------------------------------------------------------
2210       FOR I=U1 TO O1 STEP St
2220       B1=-1/Om/I
2230       G1=-B1/Q0*SQR(B1/B0)
2240       CALL Refl(G4,B4,G1,B1,Rb,C1,C1min,C1max,Rc1,Rsoll,Om)
2250       IF Rb>Minr THEN 2290
2260       Minr=Rb
2270       Lminr=I
2280       Minc=C1
2290       NEXT I
2300  ! ---------------------------------------------------------------------------
2310    MOVE I*1E6,Rb
2320    LABEL "O"
2330    GOTO 2080
2340  !
2350  ! ************* calculation of impedances and admittances *****************
2360  !
2370    C1=Minc
2380    L=Lminr
2390    Rb=Minr
2400    X23=-1/Om/C1
2410    B1=-1/Om/L
2420    PRINT LIN(3)
2430    PRINT "------------------------------------------------------------------
--------------"
2440    FIXED 2
2450    G1=-B1/Q0*SQR(B1/B0)
```

```
2460     G3=G4+G1
2470     B3=B4+B1
2480     CALL Div(1,0,G3,B3,R3,X3)
2490     R2=R3+Rc1
2500     X2=X3+X23
2510     CALL Div(1,0,R2,X2,G2,B2)
2520     P1=P0*(1-Rb^2)                      ! power delivered to the coupler
2530     Swr=(1+Rb)/(1-Rb)                   ! standing wave ratio
2540   ! ************ currents, voltages, and powers ****************************
2550     PRINT " input reflection coefficient:              r = ";Rb
2560     PRINT " VSWR:                                      s = ";Swr
2570     I2=SQR(2*P1/R2)
2580     U22=SQR(2*P1/G2)
2590     U23=I2*ABS(X23)
2600     P23=I2^2/2*ABS(X23)
2610     Pvc1=I2^2/2*Rc1
2620     P2=P1-Pvc1
2630     I3=SQR(2*P2/R3)
2640     U33=SQR(2*P2/G3)
2650     P33=U33^2/2*ABS(B1)
2660     Pv1=U33^2/2*G1
2670     I1=U33*ABS(B1)
2680     P3=P1-Pv1
2690     I4=SQR(2*P3/R4)
2700     U44=SQR(2*P3/G4)
2710     U45=I4*ABS(X45)
2720     P45=I4^2/2*ABS(X45)
2730     Pvc2=I4^2/2*Rc2
2740     P4=P3-Pvc2
2750     U55=SQR(2*P4/G5)
2760     I5=SQR(2*P4/R5)
2770     Pvc5=I5^2/2*Rc5
2780     P5=P4-Pvc5
2790     I6=SQR(2*P5/R6)
2800     Pvc=I6^2/2*Rvc
2810     Pvc6=I6^2/2*Rc6
2820     Pv=I6^2/2*Rv
2830     Ps=I6^2/2*Rs
2840     ON S2 GOTO 2850,2870
2850     PRINT "capacitor C2:                       short-circuited"
2860     GOTO 2880
2870     PRINT "capacitor C2:                       C2 = 200 [pF]"
2880     PRINT "inductance L:                        L = ";L*1E6;" [uH]"
2890     PRINT "capacitor C1:                       C1 = ";C1*1E12;" [pF]"
2900     PRINT "---------------------------------------------------------------
---------------"
2910     PRINT " impedances:     Z2: R2 = ";R2;" [Ohm];  X2 = ";X2;" [Ohm]"
2920     PRINT "                 Z3: R3 = ";R3;" [Ohm];  X3 = ";X3;" [Ohm]"
2930     PRINT "                 Z4: R4 = ";R4;" [Ohm];  X4 = ";X4;" [Ohm]"
2940     PRINT "                 Z5: R5 = ";R5;" [Ohm];  X5 = ";X5;" [Ohm]"
2950     PRINT "                 Z6: R6 = ";R6;" [Ohm];  X6 = ";X6;" [Ohm]"
2960     PRINT "---------------------------------------------------------------
```

```
---------------"
2970    PRINT " admittances:     Y2: G2 = ";G2*1000;" [mS];   B2 = ";B2*1000;"
 [mS]"
2980    PRINT "                  Y3: G3 = ";G3*1000;" [mS];   B3 = ";B3*1000;"
 [mS]"
2990    PRINT "                  Y4: G4 = ";G4*1000;" [mS];   B4 = ";B4*1000;"
 [mS]"
3000    PRINT "                  Y5: G5 = ";G5*1000;" [mS];   B5 = ";B5*1000;"
 [mS]"
3010    PRINT "                  Y6: G6 = ";G6*1000;" [mS];   B6 = ";B6*1000;"
 [mS]"
3020    PRINT "-----------------------------------------------------------------
---------------"
3030    PRINT " antenna:         Imax = ";I6/COS(Be*A);" [A]"
3040    PRINT " feeder line:     fictitious length l0 = ";L0;" [m]"
3050    PRINT "                  Imax = ";I6/COS(Bee*L0);" [A];  Ort: lmax = ";P
I/Bee-L0;" [m]"
3060    PRINT "                  Umax = ";I6/COS(Bee*L0)*Z1;" [V]"
3070    PRINT "-----------------------------------------------------------------
---------------"
3080    PRINT " available generator power:      Pa = ";P0;" [Watt]"
3090    PRINT " power, delivered to coupler:    P1 = ";P1;" [Watt]"
3100    PRINT "-----------------------------------------------------------------
---------------"
3110    PRINT " currents:                       I2 = ";I2;" [A]"
3120    PRINT "                                 I3 = ";I3;" [A]"
3130    PRINT "                                 I4 = ";I4;" [A]"
3140    PRINT "                                 I5 = ";I5;" [A]"
3150    PRINT "                                 I6 = ";I6;" [A]"
3160    PRINT "                                I33 = ";I1;" [A]"
3170    PRINT "-----------------------------------------------------------------
---------------"
3180    PRINT " voltages:                      U22 = ";U22;" [V]"
3190    PRINT "                                U23 = ";U23;" [V]"
3200    PRINT "                                U33 = ";U33;" [V]"
3210    PRINT "                                U44 = ";U44;" [V]"
3220    PRINT "                                U45 = ";U45;" [V]"
3230    PRINT "                                U55 = ";U55;" [V]"
3240    PRINT "                                U66 = ";I6*Xa;" [V]"
3250    PRINT "-----------------------------------------------------------------
---------------"
3260    PRINT " reactive powers:             P(C1) = ";P23;" [W]"
3270    PRINT "                               P(L) = ";P33;" [W]"
3280    PRINT "                              P(C2) = ";P45;" [W]"
3290    PRINT "                              P(Xa) = ";I6^2/2*Xa;" [W]"
3300    PRINT "-----------------------------------------------------------------
---------------"
3310    PRINT " dissipative powers:          Pv(L) = ";Pvl;" [W]"
3320    PRINT "                           Pv(line) = ";Pvc;" [W]"
3330    PRINT "                            Pv(rad) = ";Pv;" [W]"
3340    PRINT "                            Pv (C1) = ";Pvc1;" [W]"
3350    PRINT "                            Pv (C2) = ";Pvc2;" [W]"
```

```
3360     PRINT "                   feeder-antenna       Pv = ";Pvc6;" [W]"
3370     PRINT "                   coupler-feeder       Pv = ";Pvc5;" [W]"
3380     PRINT "-------------------------------------------------------------------
-------------"
3390     PRINT " finally radiated power:           Ps = ";Ps;" [W]"
3400     PRINT "-------------------------------------------------------------------
-------------"
3410     PRINT LIN(2)
3420     Durchl=Durchl+1
3430     NEXT S2
3440  !
3450  ! *****************************************************************
*
3460  !
3470  !
3480  !
3490  IF Durchl=0 THEN 3530
3500  EXIT GRAPHICS
3510  DISP "                   R E A D Y                    "
3520  GOTO 3590
3530  PRINTER IS 16
3540  EXIT GRAPHICS
3550  PRINT LIN(5)
3560  PRINT "                N O T   T U N A B L E                    "
3570  PRINTER IS 0
3580  PRINT "antenna not tunable"
3590  END
3600  !
3610  !
3620  !
3630  !
3640  !
3650  ! ****************************  S U B P R O G R A M S  *****************
3660  !
3670  !
3680  !
3690  !
3700  !
3710  ! **************************** complex multiplication *******************
3720  !
3730  SUB Mul(Ar,Ai,Br,Bi,Cr,Ci)
3740  Cr=Ar*Br-Ai*Bi
3750  Ci=Ar*Bi+Ai*Br
3760  SUBEND
3770  !
3780  !
3790  ! **************************** complex division **************************
3800  !
3810  SUB Div(Ar,Ai,Br,Bi,Cr,Ci)
3820  N=Br^2+Bi^2
3830  Cr=(Ar*Br+Ai*Bi)/N
3840  Ci=(-Ar*Bi+Ai*Br)/N
```

```
3850  SUBEND
3860  !
3870  !
3880  ! ***************************** reflection coefficient of coupler *********
3890  !
3900  SUB Refl(G4,B4,Gl,Bl,Rb,Cl,Clmin,Clmax,Rcl,Rsoll,Om)
3910  G3=G4+Gl
3920  B3=B4+Bl
3930  CALL Div(1,0,G3,B3,R3,X3)
3940  X23=-X3
3950  Cl=-1/Om/X23
3960  IF (Cl>=Clmin) AND (Cl<=Clmax) THEN 4010
3970  IF Cl>Clmax THEN 4000
3980  Cl=Clmin
3990  GOTO 4010
4000  Cl=Clmax
4010  X23=-1/Om/Cl
4020  R2=R3+Rcl
4030  X2=X3+X23
4040  Rb=(R2-Rsoll)^2+X2^2
4050  Rb=SQR(Rb/((R2+Rsoll)^2+X2^2))
4060  SUBEND
4070  ! ***********************************************************************
```

```
10    ! ****************************************************************
20    ! *                                                              *
30    ! *      CALCULATION OF THE TUNED ANTENNA'S BANDWIDTH            *
40    ! *                                                              *
50    ! *                     (ANTENNA TYPE 1)                         *
60    ! *                                                              *
70    ! ****************************************************************
80    !
90    !
100   !
110   PRINTER IS 16
120   STANDARD
130   PRINT PAGE
140   OPTION BASE 1
150   DIM L(101),R(101),C(101),Eta(2,82),Rf(2,82)
160   ! ******************************************* antenna data ******
170   Clmin=1.2E-11
180   Clmax=5.00E-10
190   C22=2.00E-10
200   Lmin=1E-6
210   Lmax=5.0E-5
220   Scu=5.8E7
230   Sal=3.0E7
240   Hmax=.23
250   A=7.5
260   C=3E8
270   MO=4*PI*1E-7
280   Er=1.2
290   Zl=50
300   Da=1.7E-2
310   Di=Da*EXP(-Zl/60)
320   Lt=A
330   Rq=50
340   PO=100
350   A2=1.5
360   Rc1=2.0E-2
370   Rc2=2.0E-2
380   Rc5=2.0E-2
390   Rc6=2.0E-2
400   Rsoll=Rq/4
410   RAD
420   Ql=300
430   D=7.5E-2
440   ! --------------------------------------------------------------
450   FOR Fl=3 TO 5                ! centre frequency [MHz]
460   Kont=0
470   ! --------------------------------------------------------------
480   FOR If=0 TO 81               ! frequency deviation +/- 10 kHz
490   Df=If-41
500   Kont=Kont+1
510   F=Fl*1E6+Df*1000/4*(If>0)
520   Om=2*PI*F
```

```
530   Q0=Q1*SQR(F/3E6)
540   B0=-1/Om/Lmax
550   ! -----------------------------------------------------------------------
560   H=SQR(Hmax*(Hmax-D))         ! antenna impedance
570   H1=Hmax-D/2
580   La=C/F
590   Be=2*PI/La
600   P=Be*A
610   W=2*P
620   Rv=A/D/6*SQR(F*M0/PI/Sal)*H1/H*(W+SIN(W))/W/COS(P)^2
630   Za=60*LOG(2*(H1+H)/D)
640   Xa=Za*TAN(P)
650   Z0=120*PI
660   Rs=4/3*PI*Z0*(H*F*TAN(Be*A)/C)^2
670   R6=Rs+Rv
680   X6=Xa
690   CALL Div(1,0,R6,X6,G6,B6)
700   Bp=TAN(Be*A2)/Za
710   B6=B6+Bp
720   CALL Div(1,0,G6,B6,R6,X6)
730   R5=R6+Rc5
740   X5=X6
750   CALL Div(1,0,R5,X5,G5,B5)
760   ! -----------------------------------------------------------------------
770   Durchl=0                     ! coupler tuning to centre frequency
780   IF Kont>1 THEN GOTO 1600
790   S2=2
800   Ps=0
810   ON S2 GOTO 820,840
820   X45=0
830   GOTO 850
840   X45=-1/Om/C22
850   R4=R5+Rc2
860   X4=X5+X45
870   CALL Div(1,0,R4,X4,G4,B4)
880   Z=0
890   FOR I=0 TO 100
900   L=Lmin+I*(Lmax-Lmin)/100
910   B1=-1/Om/L
920   G1=-B1/Q0*SQR(B1/B0)
930   CALL Refl(G4,B4,G1,B1,Rb,C1,C1min,C1max,Rc1,Rsoll,Om)
940   IF Rb>.995 THEN 990
950   Z=Z+1
960   R(Z)=Rb
970   L(Z)=L
980   C(Z)=C1
990   NEXT I
1000  PLOTTER IS "GRAPHICS"
1010  GCLEAR
1020  GRAPHICS
1030  SCALE -5,59,-.09,1.1
1040  AXES 10,.1
```

```
1050  LORG 5
1060  MOVE 3,.8
1070  LABEL "|r|"
1080  MOVE 45,.08
1090  LABEL "L/uH"
1100  FOR I=1 TO Z
1110  MOVE L(I)*1E6,R(I)
1120  LABEL "+"
1130  NEXT I
1140  LINE TYPE 4
1150  MOVE 50,0
1160  DRAW 50,1.1
1170  MOVE 0,1
1180  DRAW 60,1
1190  MOVE 10,0
1200  DRAW 10,1.1
1210  LINE TYPE 1
1220  MOVE 40,1.05
1230  ON S2 GOTO 1240,1260
1240  LABEL "C2: KS"
1250  GOTO 1270
1260  LABEL "C2 = 200 pF"
1270  Minr=1
1280  IF Z=0 THEN 2160
1290  FOR I=1 TO Z
1300  IF R(I)>Minr THEN 1340
1310  Minr=R(I)
1320  Lminr=L(I)
1330  Minc=C(I)
1340  NEXT I
1350  Step=0
1360  IF Minr<.001 THEN 1580
1370  Step=Step+1
1380  IF Step>40 THEN 2120
1390  IF (Step=40) AND (Minr<=1/3) THEN 1580
1400  K1=2E-6
1410  U1=Lminr-K1/Step^2
1420  O1=Lminr+K1/Step^2
1430  IF U1>Lmin THEN GOTO 1450
1440  U1=Lmin
1450  St=2*K1/Step^2
1460  FOR l=U1 TO O1 STEP St
1470  B1=-1/Om/l
1480  G1=-B1/Q0*SQR(B1/B0)
1490  CALL Refl(G4,B4,G1,B1,Rb,C1,C1min,C1max,Rc1,Rsoll,Om)
1500  IF Rb>Minr THEN 1540
1510  Minr=Rb
1520  Lminr=l
1530  Minc=C1
1540  NEXT l
1550  MOVE Lminr*1E6,Minr
1560  LABEL "O"
```

```
1570  GOTO 1360
1580  MOVE Lminr*1E6,Minr
1590  LABEL "O"                              ! end opf tuning
1600  ! ------------------------------------------------------------
1610  C1=Minc                                ! impedances and admittances
1620  L=Lminr
1630  X23=-1/Om/C1
1640  B1=-1/Om/L
1650  G1=-B1/Q0*SQR(R1/R0)
1660  R4=R5+Rc2
1670  X45=-1/Om/C22
1680  X4=X5+X45
1690  CALL Div(1,0,R4,X4,G4,B4)
1700  G3=G4+G1
1710  B3=B4+B1
1720  CALL Div(1,0,G3,B3,R3,X3)
1730  R2=R3+Rc1
1740  X2=X3+X23
1750  CALL Div(1,0,R2,X2,G2,B2)
1760  CALL Div(R2-Rsoll,X2,R2+Rsoll,X2,Rbr,Rbi)
1770  Rb=SQR(Rbr^2+Rbi^2)
1780  IF If<1 THEN GOTO 1800
1790  Rf(2,If)=Rb
1800  P1=P0*(1-Rb^2)
1810  Swr=(1+Rb)/(1-Rb)
1820  ! ------------------------------------------------------------
1830  I2=SQR(2*P1/R2)                        ! currents, voltages and powers
1840  U22=SQR(2*P1/G2)
1850  U23=I2*ABS(X23)
1860  P23=I2^2/2*ABS(X23)
1870  Pvc1=I2^2/2*Rc1
1880  P2=P1-Pvc1
1890  I3=SQR(2*P2/R3)
1900  U33=SQR(2*P2/G3)
1910  P33=U33^2/2*ABS(B1)
1920  Pv1=U33^2/2*G1
1930  I1=U33*ABS(B1)
1940  P3=P1-Pv1
1950  I4=SQR(2*P3/R4)
1960  U44=SQR(2*P3/G4)
1970  U45=I4*ABS(X45)
1980  P45=I4^2/2*ABS(X45)
1990  Pvc2=I4^2/2*Rc2
2000  P4=P3-Pvc2
2010  U55=SQR(2*P4/G5)
2020  I5=SQR(2*P4/R5)
2030  Pvc5=I5^2/2*Rc5
2040  P5=P4-Pvc5
2050  I6=SQR(2*P5/R6)
2060  Pv=I6^2/2*Rv
2070  Ps=I6^2/2*Rs
2080  IF If<1 THEN GOTO 2130
```

```
2090  EXIT GRAPHICS
2100  Eta(S2,If)=Ps
2110  Durchl=Durchl+1
2120  Eta(S2,If)=Ps
2130  PRINTER IS 16
2140  PRINT "f = ";F/1E6;"MHz    Eta =";Ps;"%     SWR=";Swr
2150  PRINT "--------------------------------------------------------
------------"
2160  NEXT If
2170  PLOTTER IS "GRAPHICS"               ! plot results
2180  GCLEAR
2190  GRAPHICS
2200  SCALE -10.5,10.5,-.06,1.15
2210  AXES 1,.1,0,0,5,5
2220  CSIZE 2.5
2230  LORG 6
2240  FOR I=-10 TO 10 STEP 2
2250  IF I=0 THEN 2280
2260  MOVE I,-2.0E-2
2270  LABEL I
2280  NEXT I
2290  LORG 4
2300  MOVE 7,2.0E-2
2310  LABEL "--> f/kHz"
2320  LORG 8
2330  FOR I=.2 TO 1 STEP .2
2340  MOVE -.4,I
2350  LABEL I
2360  NEXT I
2370  LORG 2
2380  MOVE .5,1.05
2390  LABEL "P/Pmax"
2400  MOVE 5,1.05
2410  LABEL "f = ";F1;" MHz"
2420  LORG 5
2430  FOR J=2 TO 2
2440  Ref=Eta(2,41)
2450  MOVE -10,Eta(2,1)/Ref
2460  FOR I=-10 TO 10 STEP .25
2470  DRAW I,Eta(J,4*I+41)/Ref
2480  NEXT I
2490  MOVE -10,Rf(2,1)
2500  LINE TYPE 2
2510  FOR I=-10 TO 10 STEP .25
2520  DRAW I,Rf(2,4*I+41)
2530  NEXT I
2540  NEXT J
2550  DUMP GRAPHICS
2560  PRINTER IS 0
2570  PRINT LIN(5)
2580  PRINT "Antenna type-1: Eta(f)"
2590  PRINT "a = ";A;" m; A1 = ";A2;" m    d = ";D*100;" cm"
```

```
2600  PRINT "Pmax = ";Ref;" Watt  (Pa=100W)"
2610  PRINT LIN(5)
2620  NEXT Fl
2630  END
2640  ! -------------------------------------------------------------
2650  SUB Mul(Ar,Ai,Br,Bi,Cr,Ci)          ! subroutines
2660  Cr=Ar*Br-Ai*Bi
2670  Ci=Ar*Bi+Ai*Br
2680  SUBEND
2690  SUB Div(Ar,Ai,Br,Bi,Cr,Ci)
2700  N=Br^2+Bi^2
2710  Cr=(Ar*Br+Ai*Bi)/N
2720  Ci=(-Ar*Bi+Ai*Br)/N
2730  SUBEND
2740  SUB Refl(G4,B4,Gl,Bl,Rb,Cl,Clmin,Clmax,Rcl,Rsoll,Om)
2750  G3=G4+Gl
2760  B3=B4+Bl
2770  CALL Div(1,0,G3,B3,R3,X3)
2780  X23=-X3
2790  Cl=-1/Om/X23
2800  IF (Cl>=Clmin) AND (Cl<=Clmax) THEN 2850
2810  IF Cl>Clmax THEN 2840
2820  Cl=Clmin
2830  GOTO 2850
2840  Cl=Clmax
2850  X23=-1/Om/Cl
2860  R2=R3+Rcl
2870  X2=X3+X23
2880  Rb=(R2-Rsoll)^2+X2^2
2890  Rb=SQR(Rb/((R2+Rsoll)^2+X2^2))
2900  SUBEND
```

END

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